

# Proceedings



of the

I · R · E

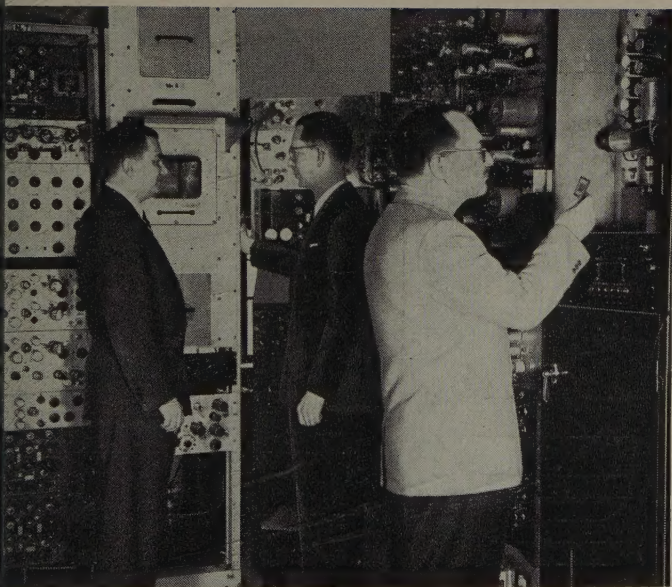
**A Journal of Communications and Electronic Engineering**

**Illinois U Library**

**October, 1950**

**Volume 38**

**Number 10**



*Hazeltine Electronics Corp.*

## **STUDIES OF NEW CIRCUITS**

Laboratories at Little Neck, L.I., N.Y., transmitting, receiving, and timing equipment (using nearly 600 tubes) enable systematic studies of circuits and equipment.

## **PROCEEDINGS OF THE I.R.E.**

Professional Groups in the IRE  
Electronics and the Electrostatic Generator  
Electronic Miniaturization  
Interference Caused by More than One Signal  
The *p*-Germanium Transistor  
Asymmetrically Driven Antennas  
Detection of Periodic Signals in Noise  
Traveling-Wave Cathode-Ray Tube  
Conductivity Measurements at Microwave Frequencies  
Cascade-Connected Attenuators (Abstract)  
Frequency Compensating Matching Sections  
S/N Improvement Through Integration in a Storage Tube  
Distortion in Angular Modulation  
Pulse Transients in Transmission Lines  
Network for Measurement of Crystal Characteristics  
Abstracts and References

TABLE OF CONTENTS, INDICATED BY BLACK-AND-WHITE MARGIN, FOLLOWS PAGE 32A

1951 Convention Authors, see page 1224

# The Institute of Radio Engineers

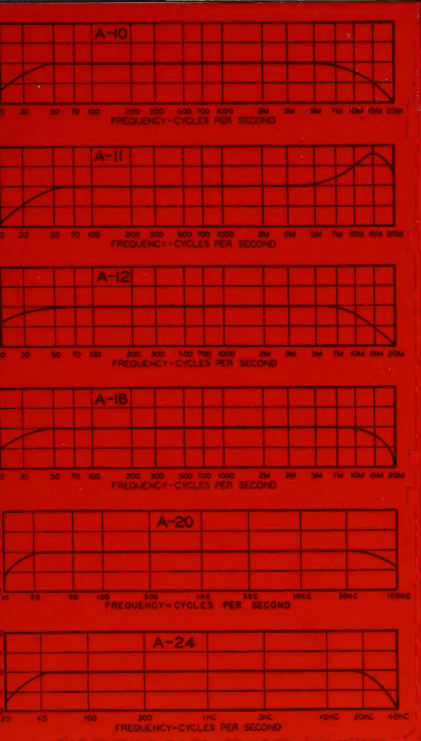


# ULTRA COMPACT UNITS...OUNCER UNITS

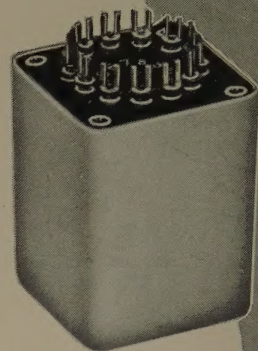
## HIGH FIDELITY... SMALL SIZE... FROM STOCK

UTC Ultra compact audio units are small and light in weight, ideally suited to remote amplifier and similar compact equipment. High fidelity is obtainable in all individual units, the frequency response being  $\pm 2$  DB from 30 to 20,000 cycles.

True hum balancing coil structure combined with a high conductivity die cast outer case, effects good inductive shielding.



Type No.	Application	Primary Impedance	Secondary Impedance	List Price
A-10	Low impedance mike, pickup, or multiple line to grid	50, 125/150, 200/250, 333, 500/600 ohms	50,000 ohms	\$15.00
A-11	Low impedance mike, pickup, or line to 1 or 2 grids (multiple alloy shields for low hum pickup)	50, 200, 500	50,000 ohms	16.00
A-12	Low impedance mike, pickup, or multiple line to grids	50, 125/150, 200/250, 333, 500/600 ohms	80,000 ohms overall, in two sections	15.00
A-14	Dynamic microphone to one or two grids	30 ohms	50,000 ohms overall, in two sections	14.00
A-20	Mixing, mike, pickup, or multiple line to line	50, 125/150, 200/250, 333, 500/600 ohms	50, 125/150, 200/250, 333, 500/600 ohms	15.00
A-21	mixing, low impedance mike, pickup, or line to line (multiple alloy shields for low hum pickup)	50, 200/250, 500/600	50, 200/250, 500/600	16.00
A-16	Single plate to single grid	15,000 ohms	60,000 ohms, 2:1 ratio	13.00
A-17	Single plate to single grid 8 MA unbalanced D.C.	As above	As above	15.00
A-18	Single plate to two grids. Split primary	15,000 ohms	80,000 ohms overall, 2.3:1 turn ratio	14.00
A-19	Single plate to two grids 8 MA unbalanced D.C.	15,000 ohms	80,000 ohms overall, 2.3:1 turn ratio	18.00
A-24	Single plate to multiple line	15,000 ohms	50, 125/150, 200/250, 333, 500/600 ohms	15.00
A-25	Single plate to multiple line 8 MA unbalanced D.C.	15,000 ohms	50, 125/150, 200/250, 333, 500/600 ohms	14.00
A-26	Push pull low level plates to multiple line	30,000 ohms plate to plate	50, 125/150, 200/250, 333, 500/600 ohms	15.00
A-27	Crystal microphone to multiple line	100,000 ohms	50, 125/150, 200/250, 333, 500/600 ohms	15.00
A-30	Audio choke, 250 henrys @ 5 MA 6000 ohms D.C., 65 henrys @ 10 MA 1500 ohms D.C.			10.00
A-32	Filter choke 60 henrys @ 15 MA 2000 ohms D.C., 15 henrys @ 30 MA 500 ohms D.C.			9.00



TYPE A CASE  
1 1/2" x 1 1/2" x 2" high

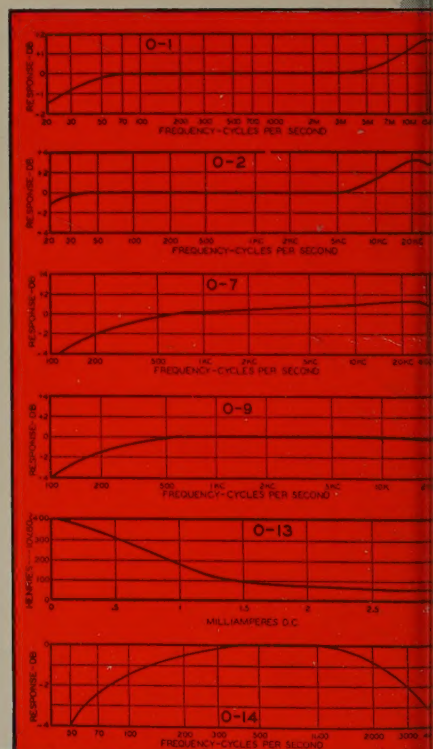
UTC OUNCER components represent the acme in compact quality transformers. These units, which weigh one ounce, are fully impregnated and sealed in a drawn aluminum housing 7/8" diameter...mounting opposite terminal board. High fidelity characteristics are provided, uniform from 40 to 15,000 cycles, except for O-14, O-15, and units carrying DC which are intended for voice frequencies from 150 to 4,000 cycles. Maximum level 0 DB.



OUNCER CASE

7/8" Dia. x 1 1/8" high

Type No.	Application	Pri. Imp.	Sec. Imp.	List Price
O-1	Mike, pickup or line to 1 grid	50, 200/250 500/600	50,000	\$13.25
O-2	Mike, pickup or line to 2 grids	50, 200/250 500/600	50,000	13.25
O-3	Dynamic mike to 1 grid	7.5/30	50,000	12.00
O-4	Single plate to 1 grid	15,000	60,000	10.50
O-5	Plate to grid, D.C. in Pri.	15,000	60,000	10.50
O-6	Single plate to 2 grids	15,000	95,000	12.00
O-7	Plate to 2 grids, D.C. in Pri.	15,000	95,000	12.00
O-8	Single plate to line	15,000	50, 200/250, 500/600	13.25
O-9	Plate to line, D.C. in Pri.	15,000	50, 200/250, 500/600	13.25
O-10	Push pull plates to line	30,000 ohms plate to plate	50, 200/250, 500/600	13.25
O-11	Crystal mike to line	50,000	50, 200/250, 500/600	13.25
O-12	Mixing and matching	50, 200/250	50, 200/250, 500/600	12.00
O-13	Reactor, 300 Hys.—no D.C.; 50 Hys.—3 MA. D.C.,		6000 ohms	9.50
O-14	50:1 mike or line to grid	200	1/2 megohm	13.25
O-15	10:1 single plate to grid	15,000	1 megohm	13.25



*United Transformer Co.*  
150 VARICK STREET • NEW YORK 13, N. Y.  
EXPORT DIVISION: 13 EAST 40th STREET, NEW YORK 16, N. Y. CABLES: "UTRARI"

Raymond F. Guy  
*President*

R. A. Watson-Watt  
*Vice-President*

D. B. Sinclair  
*Treasurer*

Haraden Pratt  
*Secretary*

Alfred N. Goldsmith  
*Editor*

B. E. Shackelford  
*Senior Past President*

Stuart L. Bailey  
*Junior Past President*

1950

Ben Akerman (6)

W. R. G. Baker

T. H. Clark

J. V. L. Hogan (2)

T. A. Hunter

H. E. Kranz (4)

F. H. R. Pounsett (8)

J. E. Shepherd

J. A. Stratton

1950-1951

A. V. Eastman (7)

W. L. Everitt

D. G. Fink

F. Hamburger Jr. (3)

H. J. Reich (1)

J. D. Reid (5)

1950-1952

W. R. Hewlett

J. W. McRae

Harold R. Zeamans  
*General Counsel*

George W. Bailey  
*Executive Secretary*

Laurence G. Cumming  
*Technical Secretary*

Changes of address (with advance notice of fifteen days) and communications regarding subscriptions and payments should be mailed to the Secretary of the Institute, at 450 Ahnaip St., Menasha, Wisconsin, or 1 East 79 Street, New York 21, N. Y.

All rights of republication, including translation into foreign languages, are reserved by the Institute. Abstracts of papers, with mention of their source, may be printed. Requests for republication privileges should be addressed to The Institute of Radio Engineers.

\*Numerals in parenthesis following Directors' names designate Region number.

# PROCEEDINGS OF THE I.R.E.

*Published Monthly by*

The Institute of Radio Engineers, Inc.

VOLUME 38

*October, 1950*

NUMBER 10

## PROCEEDINGS OF THE I.R.E.

Herman E. Kranz	1122
International Aspects of Radio	R. L. Smith-Rose 1123
3728. The Role of Professional Groups in the IRE	L. C. Van Atta 1124
3729. Electronics and the Electrostatic Generator	Burridge Jennings 1126
3730. New Techniques for Electronic Miniaturization	Robert L. Henry, Robert K-F Scal, and Gustave Shapiro 1139
3731. Interference Caused by More than One Signal	Raymond M. Wilmotte 1145
3732. The <i>p</i> -Germanium Transistor	W. G. Pfann and J. H. Scaff 1151
3733. Asymmetrically Driven Antennas and the Sleeve Dipole	Ronold King 1154
3734. Application of Correlation Analysis to the Detection of Periodic Signals in Noise	Y. W. Lee, T. P. Cheatham, Jr., and J. B. Wiesner 1165
3735. The Traveling-Wave Cathode-Ray Tube	K. Owaki, S. Terahata, T. Hada, and T. Nakamura 1172
3549. Correction to: "Speed of Electronic Switching Circuits" by E. M. Williams, D. F. Aldrich, and J. B. Woodford	1180
3736. Conductivity Measurements at Microwave Frequencies	A. C. Beck and R. W. Dawson 1181
3737. Cascade-Connected Attenuators	R. W. Beatty 1190
3738. The Design of Frequency-Compensating Matching Sections	V. H. Rumsey 1191
3739. Signal-to-Noise Improvement Through Integration in a Storage Tube	J. V. Harrington and T. F. Rogers 1197
3740. Distortion—Band-pass Considerations in Angular Modulation	Albert A. Gerlach 1203
3741. Pulse Transients in Exponential Transmission Lines	Edward R. Schatz and Everard M. Williams 1208
3455. Discussion on "On the Energy-Spectrum of an Almost Periodic Succession of Pulses" by G. G. Macfarlane	T. S. George and G. G. Macfarlane 1212
3742. Measurement of the Electrical Characteristics of Quartz Crystal Units by Use of a Bridged-Tee Network	Charles H. Rothauge and Ferdinand Hamburger, Jr. 1213
3562. Correction to: "The Transmission and Reception of Elliptically Polarized Waves" by George Sinclair	1216
Contributors to PROCEEDINGS OF THE I.R.E.	1217
Correspondence:	
3743. "Propagation of UHF and SHF Waves Beyond the Horizon"	Kenneth Bullington 1221
3744. "Suggestions to Technical Writers"	Benjamin Miessner 1222
3745. "The Transistor as a Reversible Amplifier"	W. G. Pfann 1222

## EDITORIAL DEPARTMENT

Alfred N. Goldsmith  
*Editor*

E. K. Gannett  
*Technical Editor*

Mary L. Potter  
*Assistant Editor*

## ADVERTISING DEPARTMENT

William C. Copp  
*Advertising Manager*

Lillian Petranek  
*Assistant Advertising Manager*

## BOARD OF EDITORS

Alfred N. Goldsmith  
*Chairman*

## PAPERS REVIEW COMMITTEE

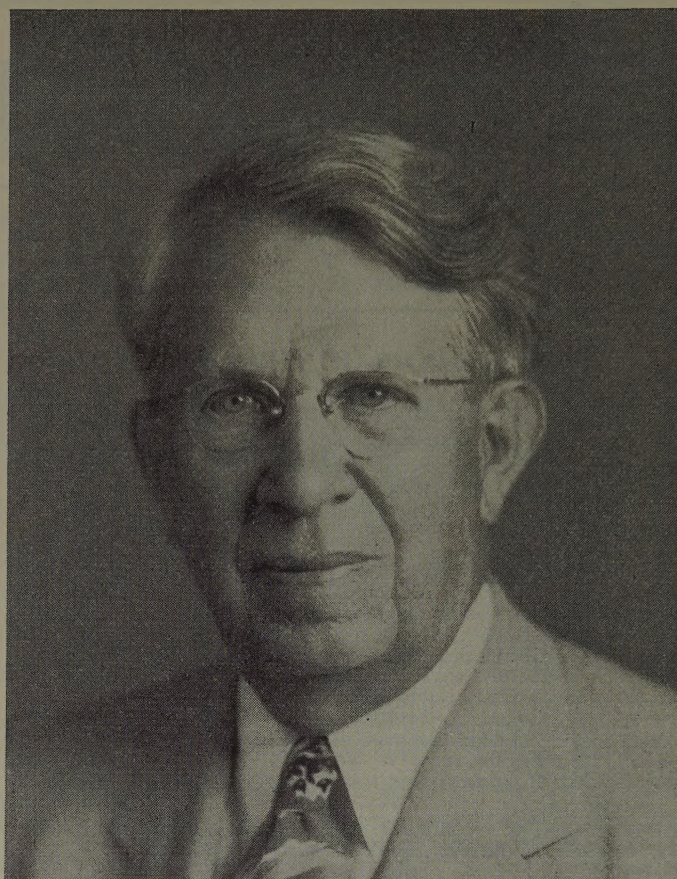
George F. Metcalf  
*Chairman*

## INSTITUTE NEWS AND RADIO NOTES SECTION

Technical Committee and Professional Group Notes	1223
IRE People	1225
Industrial Engineering Notes	1227
Books:	
3746. "The World's Radio Tubes, 1950 International Edition"	Reviewed by George D. O'Neill 1228
3747. "Traveling-Wave Tubes" by J. R. Pierce	Reviewed by Andrew Haeff 1229
3748. "Advances in Electronics, Vol. II," Edited by L. Marton	Reviewed by E. W. Herold 1229
3749. "Radio and Television Mathematics" by Bernhard Fischer	Reviewed by W. L. Behrend 1230
3750. "Television for Radiomen" by Edward M. Noll	Reviewed by Ralph R. Batcher 1230
Institute Committees—1950	1231
Institute Representatives in Colleges—1950	1233
3751. Abstracts and References	1235
3752. Books: "Applications of the Electronic Valve in Radio Receivers and Amplifiers"	Reviewed by W. C. White 1248
Section Meetings	38A News—New Products 46A
Student Branch Meetings	38A Positions Open 50A
Membership	40A Positions Wanted 56A
Advertising Index	95A

Responsibility for the contents of papers published in the PROCEEDINGS OF THE I.R.E. rests upon the authors. Statements made in papers are not binding on the Institute or its members.





## Hermann E. Kranz

REGIONAL DIRECTOR, 1950-1951

Hermann E. Kranz, Regional Director of the East Central Region, was graduated with the Bachelor of Arts degree and a summa cum laude in mathematics from Lawrence College in 1909. He entered the college of engineering at the University of Wisconsin later, where he received the B.S. degree in electrical engineering in 1914 and professional degree of electrical engineer in 1917.

He served as instructor in electrical engineering at the University of Wisconsin from 1914 to 1917. After nine years of work on manufacturing development and methods, six of which were spent at the Hawthorne Plant of the Western Electric Company, he joined the Grigsby Grunow Company as chief engineer in 1926. Under his technical guidance in 5 years this company grew from one with a daily production of a few hundred "B" battery eliminators and approximately three hundred employees to one manufacturing five thousand console radio receivers with fifteen thousand employees. He supervised facilities for the manufacture of all components which included gang condensers, rf coils, paper capacitors, tube sockets, audio and power transformers, and dynamic speakers, as well as setting up of a complete cabinet plant and a vacuum-tube plant with a capacity of seventy thousand tubes per day. In 1928 he was made plant vice-president in charge of engineering.

For nine years he served as a consultant in the radio, electronic, and automotive electrical fields. During World War II he was operations manager of the Radar Division of Research Enterprises Ltd. (a Canadian Government company) at Leaside, Ontario. One of the pioneers in the development of industrial electronic equipment, in 1933 he invented an electronic method of measuring the thickness of metallic sheet where contact could be made on one side only. He is at present the Director of the Industrial Technical Institute at Detroit.

Mr. Kranz joined The Institute of Radio Engineers as a Member in 1923 and became a Senior Member in 1947. He was one of the organizers of the Chicago Section in 1925, serving as Secretary-Treasurer from 1926 to 1928 and as Chairman, 1929-1930. He is also a past chairman of the Detroit Section, a member of its executive committee, and served for three years as a representative on the Affiliate Council of the Engineering Society of Detroit.

He is a member of the AIEE, the American Society for Engineering Education, the Radio Engineers' Club of Chicago, and the Engineering Society of Detroit. He is also a member of the Regional (III) Committee for Technical Institutes of the Engineers' Council for Professional Development.

The importance of radio communication as an agency for world-wide understanding and solidarity has been amply recognized in at least two highly diverse fashions. In the less attractive aspect, the value and influence of radio broadcasting is attested by extraordinary and large-scale efforts to interfere with broadcasts crossing the boundaries of attempted self-contained States. In the more pleasant aspect, radio is acclaimed by men of good will as a potent means for beneficial and enlightening contacts between men of all nations and modes of thought.

The following guest editorial admirably analyzes the more prepossessing facts of this situation. It has been written by a Fellow, Past Vice-President, and former Director of The Institute of Radio Engineers, who is the Director of Radio Research in the Department of Scientific and Industrial Research of the Government of Great Britain.—*The Editor.*

## International Aspects of Radio

R. L. SMITH-ROSE

The present state of modern civilization makes it quite clear that the time has long since passed when any country, state, or town could lead its own self-contained life without taking full cognizance of the activities of all its neighbors. In little more than half a century, the development of radio has reached the point at which anyone in any part of the world can be made aware of, and kept informed on, the activities of all his contemporaries distributed over the earth's surface. The radio amateur shares with the philatelist a hobby which enhances his knowledge of the world's geography and the languages and customs of other countries.

But participation in the many facets of radio science and practice can and does have a much wider influence on those engaged in it than that of merely acquiring and distributing information. It is not too much to say that the professional radio engineer and scientist has unique opportunities to know and understand his fellow human beings in other countries, and the problems which confront them in the common struggle towards a better life within our modern civilization. The installation and operation of a radio transmitter may have a very profound effect in some remote corner of the world; and the designer and operator is therefore at once forced to consider the international aspects of his action.

The demand for the unrestricted use of portions of the radio-frequency spectrum by multifarious interests made it necessary many years ago to set up an international organization to decide and control the allocation of frequencies. This has, in turn, meant that all national administrations and operating organizations have had to consider not only their own requirements, but also to try to understand the needs of others in far-off lands. While much of the discussion of the technical merits of such requirements can take place by the exchange of documents, a vital part of the whole setup consists in the arrangement of international conferences, at which the various protagonists of a case or policy meet in the flesh their contemporaries from other countries. At such meetings it is natural to find that there are many informal occasions on which the discussions and exchange of ideas are not limited to radio, or even to scientific and technical matters. International conferences provide unique opportunities for each national delegate to get to know and understand his colleagues in other countries, to become acquainted with their national habits and with the political, economic, or other problems with which they are faced. It is in this way that, of all the sciences and professions, radio plays a unique part in promoting a better understanding between peoples and nations; such understanding can do so much to avoid a common misinterpretation of the word "foreigner" and to create a spirit of unity which is today so necessary for the benefit of the world at large.

# The Role of Professional Groups in the IRE\*

L. C. VAN ATTA†, SENIOR MEMBER, IRE

Unfailing signs of the youthfulness and potentialities of any man or institution are open-mindedness toward change and flexibility of procedure. Judged from these criteria, The Institute of Radio Engineers is truly a young society, and one with its greatest accomplishments still before it.

It has become ever clearer that the membership of the Institute, now 30,000 in number, has many and diverse interests. To serve *all* of its membership adequately therefore became a matter of major importance, and one presenting certain organizational, publication, and financial problems.

To meet present and future needs, broad statesmanship has accordingly been brought to bear by the IRE Board of Directors. In orderly fashion, and without detriment to past activities or disruption of useful processes, the Institute has been expanded from a single society to what might be regarded as a close federation of a number of societies. All of these "federalized" societies work under the aegis and with the collaboration of the parent society, acting as a broad guide and helper.

This basic change is well described in the following paper by a former Chairman of one of the IRE Professional Groups and is strongly commended to the careful attention of the membership. It points out the ways in which the Institute can serve its membership with continuously increasing effectiveness and satisfaction. It also solicits vigorous co-operation between the membership and the new IRE Professional Groups.—*The Editor.*

## HISTORICAL BACKGROUND

WHEN THE IRE was formed in 1912 by the union of two local societies, radio engineers were provided with an international forum for the exchange of ideas and information. At that time engineers were broadly interested in one another's activities to a greater extent than is possible today.

In the course of the enormous growth of the radio field a great deal of specialization and divergence of interest has been inevitable. In addition to the original application of radio techniques to communication, we now have applications to navigation, radar, induction heating, and automatic computing and control. Radio techniques are combined with other techniques in such fields as acoustics, infrared, and nucleonics. Components have undergone intensive development and the operating frequency range has been greatly extended.

The Institute of Radio Engineers has grown up with the radio field to a present membership approaching 30,000. On purely quantitative grounds, any complete exchange among so many engineers is impossible. It is likewise difficult for any single engineer to encompass mentally, let alone contribute actively to the entire radio field.

Under these conditions it is natural for engineers to form, in addition, smaller, more compact groups on the basis of professional interest. This is illustrated in one field by the formation of the Acoustical Society of America some time ago, and of the Audio Engineering Society more recently. These societies attempt to meet the needs of their limited membership by specialized action.

Along with the professional advantages, which the smaller society plans to offer the individual, are attendant difficulties. The expense of the administrative staff must be carried by a smaller membership. The smaller subscription to the technical journal increases the cost per copy of publication and decreases the value of advertising space.

In general, projects assume larger proportions when undertaken by the smaller society.

These difficulties were encountered in the field of physics, where a differentiation into smaller societies occurred. The attempted solution to the problem which was worked out in this case is represented by the American Institute of Physics. It provides a number of related professional societies with a common headquarters staff and a common publications organization.

The Institute of Radio Engineers is now attempting a broadly analogous solution in the field of radio, but without the painful intermediate stage of entirely separate societies. The Professional Groups within the IRE are intended to provide for specialized professional needs without losing the economy of size in meeting overhead expenses. In addition, there is the desire to preserve that important over-all unity of interest appropriate to the field of communications and electronics.

The Professional Group principle of operation was adopted by the IRE Board of Directors in March, 1948. Since then all the necessary implementing steps have been taken, as follows:

- (1) Appropriate amendments to the By-laws of the Institute have been adopted
- (2) A permanent Professional Groups Committee has been established
- (3) A model constitution and bylaws have been prepared as general guidance to individual Groups; and finally
- (4) A detailed Professional Group Manual, revised in May, 1950, has been published for the guidance of Group promoters.

The IRE has followed the wise course of limiting itself to stimulation and guidance of the Professional Group development. It has established general policy and procedure, but has left the formation and activities of individual Groups to the initiative of interested IRE members.

Up to date there have been nine Professional Groups established with the following fields of interest and memberships:

Field of Interest	Approximate Membership
Antennas and Propagation	813
Audio	1,051

Broadcast Television Receivers	849
Broadcast Transmission Systems	514
Circuit Theory	1,227
Instrumentation	1,300
Nuclear Science	456
Quality Control	234
Vehicle and Railroad Radio Communications	246

It should be emphasized that these figures represent current and incomplete returns. These Groups have not been uniformly active, nor are their problems identical. It is not expected that they will follow altogether similar courses of development. Several additional Groups are in various stages of formation.

In carrying out their objectives the Groups have engaged in such activities as the following:

- Sponsored independent national symposia
- Sponsored sessions at the RMA/IRE Radio Fall meeting in Syracuse
- Sponsored symposia at the 1950 IRE National Convention
- Formed units co-operatively with IRE Sections
- Issued periodic Newsletters
- Undertaken the reorganization and recommendation of certain papers for publication in the PROCEEDINGS
- Arranged for the consideration and recommendation of tutorial papers for the PROCEEDINGS OF THE I.R.E.

The Groups are by no means limited to these activities.

With this background we may inquire regarding the relation of the Professional Groups to the IRE, to the individual, to the local Section, to the IRE Technical Committees, and to other societies. In considering these relations we shall obtain a clearer picture of the role of the Professional Groups.

## RELATION TO THE IRE

The Professional Group is a subdivision of the membership of the IRE based on professional interests and needs, and established under a Professional Group constitution. The constitution defines the technical

\* Decimal classification: R060. Original manuscript received by the Institute, June 24, 1950. Presented, IRE/URSI Antennas and Propagation Symposium, San Diego, Calif., April 3-5, 1950.

† Hughes Aircraft Company, Culver City, Calif.

field of interest of the Group, establishes an administrative committee and such supporting committees as are deemed necessary, prescribes broadly the functions and procedures, and fixes a minimum level of activity.

The Administrative Committee of the Professional Group reports periodically to the Professional Groups Committee of the IRE and depends upon it for general guidance as to policy and procedure. The Group requests of the IRE financial assistance as required, and turns over net receipts to the Institute. The Group depends upon the Headquarters Staff for maintaining mailing lists, for printing and distributing material as requested to the Group membership, for co-operative papers procurement, and for printing notices in the PROCEEDINGS. The PROCEEDINGS is the only outlet of the Group for technical publications and has the responsibility of providing adequate space for suitable publications in the field of interest of the Group, within the financial limitations of the IRE.

The Group provides the IRE with the means for decentralized control in the planning of national symposia in its field and in the arrangement of specialized sessions at Conventions. It provides the IRE with a national framework for the co-ordination of sectional activities in its field. With a view to the special needs of the field, the Group represents viewpoints or pleads cases with the IRE and proposes actions.

#### RELATION TO THE INDIVIDUAL

Of more immediate concern to the individual is the relation of the Professional Group to him. In many respects the Group serves the individual as the satisfactory equivalent of a small, select society in his own field of specialization. This is made possible by the freedom granted to the Administrative Committee in planning and conducting the activities of the Group. Thus the individual has all the advantages of membership in a large general society as well as a specialized smaller society.

The principal activity is now the arrangement of specialized symposia. These symposia may be made to coincide with IRE Conventions to reduce travel requirements and increase attendance. Or they may avoid the Conventions in order to achieve a greater concentration of papers and specialists and a more appropriate atmosphere. However, the keynote must be an intimate appreciation of the subject matter and of the needs of the individual member. Thus meetings may be conducted at a more leisurely pace, or a subject of current importance may be intensively explored.

In connection with its symposia, the Group may seek out papers covering interesting work and may later recommend publication in the PROCEEDINGS to the author and to the Editor. The Group may survey past issues of the PROCEEDINGS to determine whether its field has been properly represented in IRE publications. (The Groups have been asked to do this by the Editorial Department.) The possibility of the Group assuming formal responsibilities for the procurement and review of papers in its field for publication in the PROCEEDINGS has not yet been fully explored, but already close liaison

exists between the Groups and the Board of Editors, the Papers Review Committee, and the Editorial Administrative Committee. The Groups are urged to request, and will receive, adequate representation on each of these bodies.

The Group maintains a list of active members and distributes notices of national or sectional meetings in its field. These notices may be expanded to include a variety of current events of special interest. The Group may compile and distribute technical or personnel information. And, in general, it may serve as a source of information in response to specific inquiries from its members.

The general membership of the Group is simply a mailing list for announcements. The individual member remains merely an item on this mailing list, except as he changes this through participation in Group activities. After all, these activities are the work of individuals and depend upon their contributions. One type of responsibility of the Group member is to attend technical meetings, to participate in the discussion, and when possible, to present a paper. A second type of responsibility is for planning Group policy and activities through correspondence with, or membership on, appropriate technical committees. In many cases this committee service would start in the local units associated with IRE Sections.

A member who has recognized a real or fancied inadequacy in the IRE peculiar to a particular field now has a natural spokesman. The appropriate Professional Group can consider the extent of the inadequacy and defend recommendations for correction in the proper quarters in the IRE. As an example, it has been pointed out recently that the IRE has not a single Fellow in the field of Acoustics! An active Professional Group in this field in the past might have prevented this situation. It can now be readily corrected.

#### RELATION TO THE SECTION

We must explore the relation of the Professional Group to the IRE Sections to find its source of continuing strength. The Section respects a division of IRE activities on a geographical basis. The Professional Group is national or international in scope, but divides the field of electronics on the basis of technical subject matter.

The Sections should be informed in the near future of the status of the Professional Group movement and should be asked to indicate the extent of their interest in particular Groups. Each Group should carry on correspondence with interested Sections with a view to establishing professional units in these Sections. A less active professional unit might require only a representative on the Section Program Committee. A more active unit might require a local Planning Committee and a delegate to the Administrative Committee of the Group. The Groups should invite such delegates to attend the meetings of its Administrative Committee. These delegates would supply new ideas for expanding the activities of the Group. They would constitute an important source of new members for the Administrative Committee.

The Group publicizes through its mailing list those technical meetings in its field arranged by the Sections. It might act as a

clearing house for the scheduling of such meetings, to avoid interference with each other and with national meetings. It might assist local meetings by suggesting session topics or by locating speakers.

#### RELATION TO THE TECHNICAL COMMITTEE

The relation of the Professional Group to the IRE Technical Committee has been questioned with the thought that there may be overlap of function. However, this is not the case.

The Technical Committees have prescribed for them an exacting and important but very definite task: the standardization of definitions of terms and the initiation of techniques and methods of measurement. The work can be anticipated for years in advance, and must be repeated periodically to keep the standards up-to-date. In this task the Technical Committees are, as they must be, closely co-ordinated by the Standards Committee. The emphasis is on scholarship rather than initiative.

The Professional Group, on the other hand, has a very broad charter and a task only vaguely defined; in effect, to determine the professional needs in the field of the Group and to carry out or promote actions designed to meet those needs. The Group will stand or fall, depending on the imagination with which it approaches its problems and the energy with which it champions its causes. Its sphere of activity will probably increase as its possible functions are better understood by all concerned.

#### RELATION TO OTHER SOCIETIES

The possibility of relations between the Professional Group and other societies is established by the Group Constitution which allows for co-operative meetings with other technical organizations. These co-operative meetings have proven very successful because of the additional support of joint sponsorship. It should be recognized that the alternative would frequently be two interfering meetings.

The San Diego Symposium on Antennas and Propagation is a good example of co-operative effort. The sessions of April 3 and 4, 1950, were arranged by the Professional Group on Antennas and Propagation, were co-sponsored by Commissions 2, 3, and 6 of the USA National Committee, URSI, and were supported locally by the San Diego Section. The Armed Services arranged a technical Military Conference of four sessions on the same subject for April 6 and 7. Finally, the United States Navy Electronics Laboratory provided space and facilities for all the sessions. Actually, co-operative effort may make the difference between success and failure in a meeting of this kind.

In future meetings, we can visualize co-operation of a Professional Group with other Groups, with IRE Sections or Regions, with RMA and URSI, with the Armed Services, with units of the Research and Development Board, and with other similar groups. This co-operation in technical meetings is practicable in spite of the varied ultimate interests of the several organizations. For example, the Professional Group is concerned with providing a technical outlet for its members and proper material for the PRO-

CEEDINGS. By contrast, the URSI is concerned with the selection of material in this country for presentation at its international General Assembly. By further contrast, the Armed Services and the RDB are interested in stimulating any exchange of information which will result in more and better military equipment.

#### PROCEDURES

The common concern of those interested in taking any initiative in the Professional Group movement is necessarily with procedures: procedures for promoting a Group, for forming a professional unit in a Section, for joining an existing Group, for conducting a symposium, and for furthering the publication of papers in the PROCEEDINGS. Considerable effort has been made by the Professional Groups Committee to define procedures in the Manual. There has been some guidance from experience, but it is very easy in this work to get into new territory and to find oneself without ground rules.

The steps recommended for one who is interested in promoting a Professional Group are:

- (1) Write to the Technical Secretary of the Institute for general information, stating the field of interest of the contemplated Group.
- (2) Locate 5 to 15 others interested in promoting the Group.
- (3) Prepare a petition and obtain at least 25 signatures.
- (4) Forward the petition to the Chairman of the Professional Groups Committee through the Technical Secretary at the Institute Headquarters.
- (5) After approval of the petition by the IRE Board of Directors, forward suggested names for the initial Administrative Committee through the same channel.
- (6) After notification of appointment of the Administrative Committee, call a meet-

ing of the Committee. At this meeting elect a Chairman, prepare a Constitution and By-laws, and name officers and supporting committees.

(7) Submit the Constitution to the IRE Professional Groups Committee for approval. Report the other actions to the Technical Secretary for his information. With the approval of the Constitution, the Professional Group is in formal existence.

Procedures for forming a professional unit in a Section have not been so precisely defined. The first step would be to communicate with the Chairman of the corresponding Group, whose name could be obtained from the Technical Secretary of the Institute. The next step would be for the Section Program Committee to name one or more Group members to take care of local activities in the field of the Group. Finally, if invited to do so by the Group, the Section could name a delegate to the Group Administrative Committee meetings. In any case, active liaison should be maintained with the Chairman of the Administrative Committee.

When an IRE member of Associate grade or higher wishes to join an existing group, he writes the Group Chairman or the Technical Secretary of the Institute to this effect. He will be informed automatically that his name has been added to the membership list, unless the Group Constitution establishes special qualifications for membership.

For conducting a symposium, the Manual provides guidance that would take care of the most elaborate meeting. However, for an average symposium, reasonable requirements would be: (1) A Steering Committee consisting of representatives of the sponsoring organizations; (2) a symposium chairman; (3) a technical program committee which might consist of session chairmen responsible to the symposium chairman; (4) officers for finance, program publication, and

local arrangements responsible to the symposium chairman; and (5) officers for selecting and recommending papers for publication in the PROCEEDINGS.

This is enough to indicate that procedures have been defined or can be invented to take care of recognized Professional Group activities.

#### CONCLUSION

In conclusion, specialized professional needs within the IRE have been recognized and the Professional Group principle has been adopted to meet these needs. The Group movement is well launched and gives early promise in some areas. Progress has been made toward defining the relations of the Professional Groups to the IRE, to the individual, to the Section, to the Technical Committee, and to other societies. Detailed procedures have been established for many of the Professional Group activities.

I have dwelt at some length on each of these subjects and may have created the impression that the success of a Professional Group is guaranteed by its charter. If so, let me hasten to add some reservations. The success of a Professional Group depends upon several factors: the proper choice of field of interest, a cohesive feeling among the members of the Group, Administrative Committee members and Section representatives willing to give freely of their time and thought, and a consistently sympathetic reaction in higher IRE circles to the recommendations of the Group.

Much remains to be done in initiating and activating Groups. Their proper functions must be further explored. All this may be expected to require several years of persistent effort. I venture to predict that by the time the Professional Group principle has been fully implemented, there will have been major changes in the IRE practices, particularly with regard to meetings and some aspects and procedures of publications.

## Electronics and the Electrostatic Generator\*

BURRIDGE JENNINGS†

**Summary**—The role of the electrostatic generator as a particle accelerator in the field of nuclear physics is that of a precision instrument. It is capable of producing scientific data with high voltage accuracy and is, within its limited voltage range, an almost ideal tool for this type of research. The design of the generator, its advantages, limitations, and the auxiliary equipment necessary to make it a tool for nuclear research are discussed. Emphasis has been placed on the function of electronics in this field both in the control and measurement circuits, and in the general problem of the electrostatic generator.

#### INTRODUCTION

THE ELECTROSTATIC generator is one of the tools used in the experimental study of nuclear physics. It is

a source of high energy particles, like the cyclotron, betatron, and synchrotron, but differs from these machines in that it directly generates the total potential instead of accelerating the ions many times by the same small voltage increment. This machine, although limited in maximum obtainable voltage and ion current, is, within its range the best generator for precision nuclear physics. For this reason it is very important to understand its present design and extend its present limitations.

Electronics in the narrow sense, involving electronic vacuum tubes of fixed characteristics and circuits composed of resistors, condensers, inductances, and resonance circuits, has as yet little use in the electrostatic generator. This type of electronics is used in control and measurement problems, but these are usually minor and the circuits are

simple. If "electronics" is defined in a broader sense as the study of the interaction of electric and magnetic fields with electrons and ions, the electrostatic generator is just one big electronics problem. This paper will be largely concerned with present day solutions to this problem.

#### NUCLEAR PHYSICS

The most general problem of nuclear physics today is to understand how the heavier elements or isotopes are formed from the so-called elementary particles such as the neutron, proton, meson, neutrino, and perhaps others. It is probable that these elementary particles exist within any nucleus—at least it is possible to extract most of them by various types of natural or artificial radioactive disintegrations or transmutations. It would also seem reasonable that a knowledge

\* Decimal classification: 537.23. Original manuscript received by the Institute, June 2, 1950.

† Westinghouse Electric Corporation, Research Laboratories, East Pittsburgh, Pa.

of the interaction forces—such as the magnitude, sign, and range of the forces between various elementary particles, and a knowledge of the laws of interaction of these various forces in a nucleus should aid greatly in giving an answer to this general problem.

As none of the elementary particles or the individual nuclei formed of them are directly visible, we must study them by their physical effects using various types of instruments such as Geiger counters, scintillation counters, Wilson cloud chambers, and the like, and deduce what basic phenomena we can from these effects. These experiments are, philosophically, very similar to optical processes in which one illuminates an object to be studied with a beam of light and examines the color and intensity of the light reflected from or refracted by this object and deduces its properties from these data. In the case of nuclear studies, the object, or target, may be examined by "illuminating" it with various types of beams, of protons, neutrons, deuterons or gamma radiation, etc., and investigating the results with some indicating or detecting system. In general, two things are found to occur. Particles in the original beam may be scattered from the target material without change in type but only a change in direction and energy; or, if the energy relations permit, there may be a transmutation resulting in the production of a different emergent particle with perhaps violently different energy.

The experiments resulting in simple scattering are of great importance when both the beam and target are composed only of elementary particles, since range and magnitude of the forces between these particles may sometimes be deduced from the angular distribution of the scattered beam. This type of experiment is especially important in the case where protons are scattered from protons, thereby yielding the proton-proton forces, and where neutrons are scattered from protons giving information about the forces that act between the neutron and proton. Neutron-neutron scattering experiments would be of great interest but so far are experimentally impossible.

If we use as target material some element or isotope formed of many elementary particles, scattering of the impinging beam may occur in the same way, but in addition many other processes may take place. When the beam particle strikes the target it may be scattered by the target material or, if it has enough energy, it may enter one of the nuclei and cause a reaction resulting in the emission of a particle of another type and with a very different energy. Nuclear physicists think of such a reaction as the formation by the target nucleus and beam particle of a compound nucleus of an element which has too much energy to be stable, and breaks up later with the emission of a nuclear particle and perhaps a gamma ray. This compound nucleus represents only a very short-lived step in the reaction, but nevertheless it has been found possible to find energy levels or resonance levels in this nucleus. It is possible, in analogy with the energy levels or spectral lines found in atomic studies, that the spacing and intensity of these resonance levels may provide some clue to the structure of this nucleus.

The general experimental conditions

which are necessary for studies of this type are: a beam of particles, either protons, deuterons, neutrons or perhaps electrons, small and well collimated, and traveling in a well-defined direction. The velocity or voltage of this beam must be accurately known, very constant and sufficient to excite the desired reaction. The beam must contain enough particles to allow the results of the reaction to be measured by the presently developed radiation or particle detection systems. A target element, preferably of high isotropic purity, is also necessary, either thick or thin as the experiment dictates, and counters or a Wilson cloud chamber to measure the results of the reaction. It is important that the beam is not contaminated with spurious particles formed in the generator. Unwanted disintegrations may also be caused if a portion of the beam impinges on other materials than the desired target.

The electrostatic generator developed by Van de Graaff<sup>1</sup> and others<sup>2-5</sup> fortunately fulfills many of these requirements. The voltage by which the ion beam is accelerated appears as a true dc potential in the machine, so the velocity or energy of the ions striking the target is known with the same accuracy as this dc voltage. The spread in the beam voltage or the inhomogeneity of the beam energy is dependent on the stability of the generator voltage and very little else. The adjustment of the beam voltage over a very wide range can be very simply accomplished. The ion beam in such a generator, originating in an ion source inside the high voltage electrode, travels in the form of a focused beam until it strikes the target material, so that the background of radioactive disintegrations produced by high velocity particles striking other elements than those in the target is very low.

These conditions are not obtained in the cyclotron, linear accelerator, or other type of instrument in which the ion is accelerated many times by a small repetitive voltage. A cyclotron, although at present able to produce a voltage many times higher than an electrostatic generator and with much higher ion current, can, in general, produce only an ion beam of a single voltage dictated by its magnetic field and rf frequency. Reduction in the particle energy is usually made by inserting foil absorbers in the beam which causes a further increase in the beam voltage spread which may be excessive already because of the manner of acceleration. Only a part of the cyclotron ion current strikes the desired target; the remainder impinges on the interior of the cyclotron vacuum chamber. For this reason the cyclotron is itself a prolific source of neutrons and gamma rays of all energies, coming from all directions so that shielding is difficult. The

difficulty of doing precision experiments with a cyclotron increases rapidly when one attempts to greatly reduce the error of measurement.

The electrostatic generator in its present form is limited both in ion beam current and maximum operating voltage. The actual top operating range of existing generators is about 4 Mev, although a generator has attained 5 Mev, while accelerating electrons, and there are two generators now in the design stage whose designers have set 10 to 12 Mev as a goal. Many of the present generators in this country, built in the early days of generator design, have fallen short of their expectations by a factor of two or more. The reasons for this are not well known, although it is certainly true that the laws of voltage breakdown as experimentally determined in the region of a few hundred kilovolts cannot be extrapolated reliably to very high total voltages. Some progress has been made in reducing the size of electrostatic generators by using concentric electrode design and electro-negative gases such as sulfur-hexafluoride or freon instead of air.

#### THE ELECTROSTATIC GENERATOR

The total ion accelerating voltage in an electrostatic generator appears between an insulated electrode and ground. A belt composed of an insulating substance such as rubber or cotton is used as a conveyor to carry charge up to the electrode. Charges are placed on this belt at the ground end by a corona process and are carried up to the electrode against the repelling potential of the already charged electrode by the power of the motors driving the belt. The belt charging equipment consists of a row of sharp points, such as ordinary pins placed very close to the surface of belt, and charged to 20 or 30 kv by some type of transformer-rectifier set. Corona, formed at these sharp points, sprays charge on the rapidly moving belt just in front of them. Since the belt is an insulator, the charge cannot move freely. A set of needles of the same type removes the charge in the electrode, where it flows to the outside surface as the electrode acts as a Faraday cage. The voltage attained by the electrode is determined by the balance between the current carried up the belts and that lost by all processes to ground. A typical electrostatic generator of present design is shown in Fig. 1. The electrode is supported by a number of insulating columns of porcelain or plastic suitably protected by a potential dividing system. The electrostatic belt and ion acceleration tube are placed in the same neighborhood as the support members so that they can be protected by the same system. The belt driving motor, belt pulley, and corona spray units are at the bottom end of this column; the upper belt pulley, ion source, ion source power supply, and upper spray equipment are in the electrode.

The entire apparatus is contained in a pressure vessel capable of withstanding the desired air pressure. The high pressure air serves simply as an insulator to prevent electrical breakdown between the high voltage electrode and ground. The shell of the pressure tank as shown in Fig. 1 can be unbolted from the ground plate and lifted up for servicing.

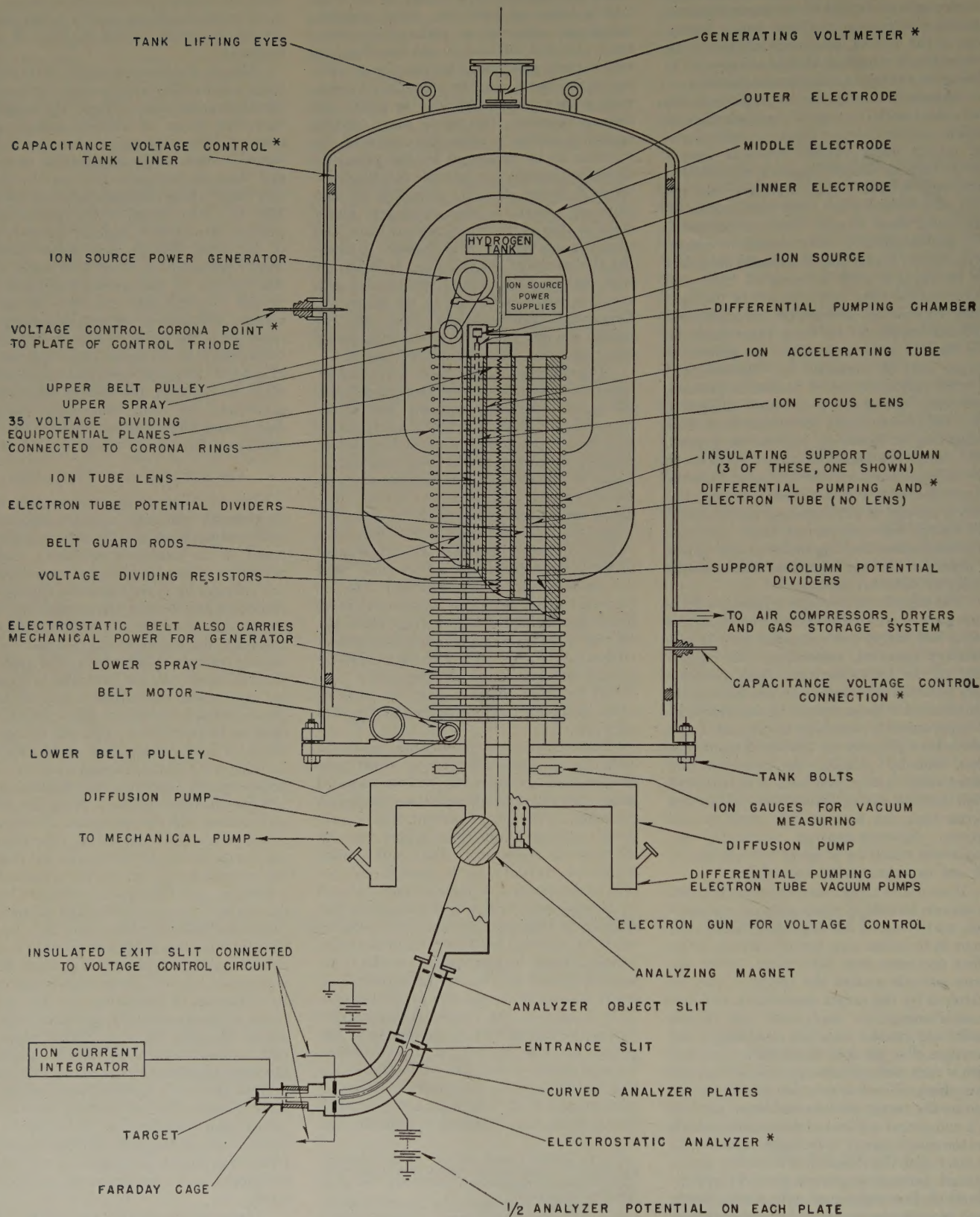
<sup>1</sup> J. G. Van de Graaff, K. T. Compton, and L. C. Van Alta, "The electrostatic production of high voltage for nuclear investigations," *Phys. Rev.*, vol. 43, p. 149; February, 1933.

<sup>2</sup> R. G. Herb, D. B. Parkinson, and D. W. Kerst, "The development and performance of an electrostatic generator operating under high air pressure," *Phys. Rev.*, vol. 51, p. 75; January, 1937.

<sup>3</sup> M. A. Tuve, L. R. Hafstad, and O. Dahl, "High voltage technique for nuclear physics studies," *Phys. Rev.*, vol. 48, p. 315; August, 1935.

<sup>4</sup> W. H. Wells, R. O. Haxby, W. E. Stephens, and W. E. Shoupp, "Design and preliminary performance tests of the Westinghouse electrostatic generator," *Phys. Rev.*, vol. 58, p. 162; July, 1940.

<sup>5</sup> For other references, see "Particle Accelerators," Brookhaven National Laboratories, BNL-L-101, pp. 17-26.



\* ITEMS SHOWN WITH STAR WOULD PROBABLY NOT SIMULTANEOUSLY BE ON ANY ONE ELECTROSTATIC GENERATOR

Fig. 1—Schematic of an electrostatic generator.

## VOLTAGE DIVIDING SYSTEM

The design of an electrostatic generator to produce the greatest peak operating voltage in a given tank size is largely a series of compromises. The high voltage electrode size is a compromise between the space required inside for the ion source and power supplies and the best estimate of the maximum allowable field intensity of the electrode surface. This field intensity may be the maximum allowed anywhere in the generator because the gas under pressure is the best insulator. The field intensity down the support columns, belts, and accelerating tube must be reduced to the lowest possible value, because the breakdown of the insulating material in this direction is not improved by gas pressure. This field intensity is minimized by making the distance between the electrode and ground along the support columns four to six times that across the air gap and by forcing the voltage distribution to be linear in this direction. This is accomplished by constructing the support insulators and ion tube of many short insulating sections about 3 to 5 inches long, with a metal conductor between each section. Such a section in the acceleration tube is shown in Fig. 2. The ion tube and insulated support members therefore consist of a series of insulated sections and equally spaced metallic conductors which are electrically connected in each plane to form an approximation of a conducting sheet completely across the support column region. The edges of each of these planes are protected by a hoop or ring of tubing around the outside to reduce the voltage gradients at this point, and the belt runs are protected in a similar manner. Each of these metallic planes across the support column region is forced by means of a resistor or corona voltage divider to take such a potential as to make the voltage roughly linear between the electrode and ground

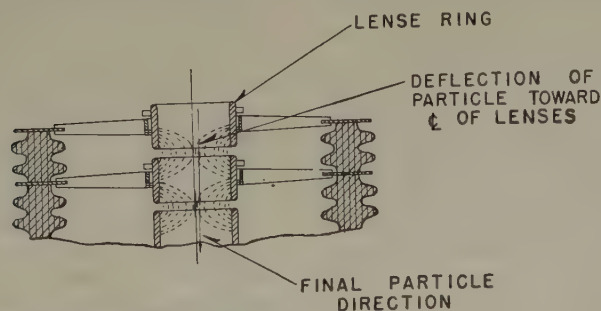


Fig. 2—Section of a vacuum tube showing electrostatic lenses.

It is very important, however, to choose the spacings between electrode and ground through the air path and down the support column such that electrical breakdown will occur first across the air path. A flashover down the support column is very likely to rip the charging belt or do other serious damage which must be absolutely avoided.

The ions are accelerated within the ion acceleration tube by a series of cylindrical electrodes, each connected to an equipotential plane of the voltage dividing system, and are focused by the lens properties of this type of electrode. Originally, simple cylindrical lens of type shown in Fig. 2 were used but these have been supplanted by the re-en-

trant design shown in Fig. 3. This change was introduced to eliminate the effect on the beam of high fields which might result from the collection of charged particles on the inside insulating surfaces of the ion tube. The re-entrant lens system shields the ion beam from such fields. This type of tube design is particularly desirable for the lenses near the high voltage electrode where the ion beam is travelling with a low velocity. The efficiency of focus lenses after the initial acceleration of the beam is questionable. Focus lenses are sometimes left out, thereby increasing the ion tube pumping speed.

Recent generator design has also made use of the principle of dividing the potential between electrode and ground (see Fig. 1) by adding concentric electrodes around the main one and connecting them to the potential divider at such a voltage as to get an advantageous field distribution around the electrode. This has many advantages, but installing and supporting these shields and servicing the inside electrode and its equipment is a difficult problem. However, such a

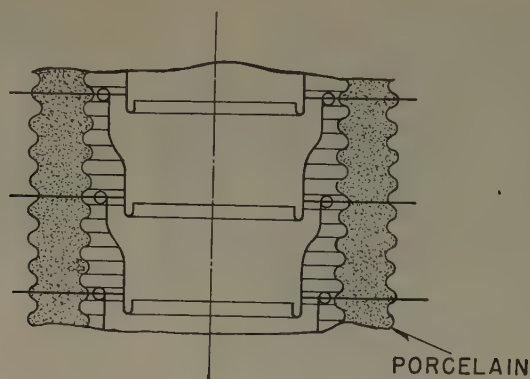


Fig. 3—Re-entrant tube design.

The use of air as an insulating gas is dictated largely by its availability. It must be dried because water vapor increases the leakage current to ground. Most electrostatic generators have large air driers of activated alumina or silica jell to reduce the humidity of the air before it is put in the tank. It has been found that the use of such electronegative gases such as Freon or sulfurhexafluoride, either pure or mixed with air, will increase the breakdown voltage by as much as a factor of 2 for a given pressure.

We encounter here the most important electronics problem of the electrostatic generator because the voltage limitation is its most serious fault. The electrical breakdown of gas under high pressure is not well understood, and its study is made more difficult because the results of small-scale experiments using low voltages and short spacings cannot be extrapolated with any accuracy to high voltages at large spacings. Although the breakdown mechanism is sufficiently well understood so that predictions can be made under ideal conditions, these conditions do not prevail in a practical design; consequently, the voltage breakdown of a generator may fall far short of that predicted from small scale measurements.

It is interesting to note, at this point, that the operating voltage of an electrostatic generator may be governed by some time dependent phenomena as the generator must operate for a very long time. It is known that impulse generators, magnetrons, and other electronic devices can operate at far higher voltages if the voltage is in the form of a very short pulse. An impulse generator can produce pulses of 5 or 6 million volts with an insulation design which would hold less than a million volts if operated as a dc generator.

Perhaps the most interesting and hopeful approach toward increasing the voltage of an electrostatic generator is in the possible use of high vacuum as insulation. Studies<sup>6</sup> are now under way to better understand the breakdown in an evacuated ion acceleration tube, but one eye is being kept on this work to see if it is applicable to vacuum insulation of the entire machine. At the present time vacuum cannot compete as an insulation with high pressure insulation in an electrostatic generator. However, relatively low

design plus the use of various electronegative gases allow voltages of the order of 3 to 5 million to be obtained in a comparatively small pressure tank.

In the more elementary generators where the insulation was simply air at atmospheric pressure, immense machines were required to give over one million volts potential. The next step in design was to enclose a large machine in a pressure vessel and increase the operating pressure to several atmospheres. Large machines were designed in this era of development, notably those at the Westinghouse Research Laboratories and at the Department of Terrestrial Magnetism in Washington, D. C.

<sup>6</sup> J. G. Trump and R. J. Van de Graaff, "The insulation of high voltages in vacuum," *Jour. Appl. Phys.*, vol. 18, p. 327; March, 1947.

voltage studies over short spacings have indicated the possibility of holding very high voltage gradients. As the spacing is increased, the limiting gradients go down, until vacuum insulation in any reasonably designed generator would not be as good as high pressure and its use is attended by very difficult and unsolved engineering problems. However, as the process of vacuum breakdown is very poorly understood, it is possible that the ability to stand high gradients may not be a function of the spacing but rather of the vacuum, surfaces, etc., which may be under our control when a better understanding is attained.

#### ION SOURCES

Another electronic problem in electrostatic generators is the production of the charged particles or ion beams suitable for accelerating by the generator. Two polarities of particles can be used: electrons, and those with a positive charge, or protons, deuterons, tritium ions and helium ions. Although not in such wide use in the nuclear field, electron accelerators are used as generators of penetrating high energy X rays. In general the techniques for producing electron beams are well worked out, and the gun-type cathode emitters such as are employed in cathode-ray tubes and electron diffraction equipment are used.

Positive ion sources are much more difficult to design, build, and use. The positive ions are formed from the gas, say hydrogen, by electron bombardment in some type of discharge tube. Of the several types of ion sources in use the most common and simplest type is that of the filament supported arc shown in Fig. 4. The filament, generally of the oxide-coated type, emits electrons which are accelerated by a potential of about 50 volts between it and an anode. These electrons collide with hydrogen gas molecules in their path and ionize the atoms in these molecules to form a proton, and, of course, a free electron. The original electron, having lost much of its energy, is again accelerated by the field and may repeat the process before it has lost one of the electrodes. The electron from the ionized atom does the same thing, so the process could easily become a chain reaction of the type found in conventional discharge and be independent of the electrons from the filament. However, because it is desirable to keep the hydrogen pressure low for other reasons, the arc is usually not operated in the self-sustaining region.

The ions formed in the glow discharge in the arc chamber are extracted from the arc by a probe. This electrode is operated at a potential of 300 to 5,000 volts negative with respect to the arc chamber, and its field, penetrating into the arc chamber a short distance through the small arc chamber hole, pulls out a small number of the ions from the discharge. Those ions which happen to go into the small probe hole are accelerated by the probe potential and come out into the acceleration tube in a fairly well collimated but still divergent beam. This diverging beam must be made to converge to a focus on the distant target at the ground end of the generator by a cylindrical "focus" lens, located a few inches below the probe exist, and supplied by an adjustable source of 10 to 20 kv.

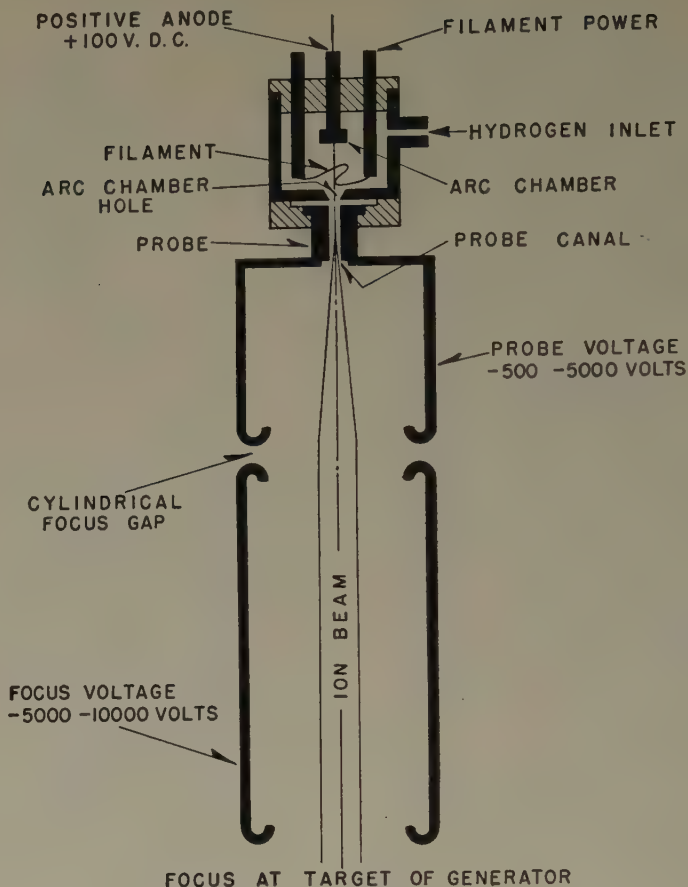


Fig. 4. Simple filament supported ion source.

More advanced designs have attempted to eliminate some of the defects in this type of ion source. The use of any filament in a hydrogen atmosphere at these low pressures is very difficult and the life of the filament may be short. Attempts to eliminate the filament have led to the use of the rf excited arc in the ion source. The arc discharge produces several types of ions; unfortunately molecular ions of the types  $HH^+$  and  $HHH^+$  are produced in larger quantities than protons. These ions are accelerated by the generator potential to the same total energy as the proton, but this energy is divided between either two or three particles so the effective energy in a nuclear reaction is only one-half or one-third of the proton energy. They are of little use in nuclear physics and design attempts are made to reduce their numbers. The production of these unwanted ions has been found to occur, at least partially, at the walls and metal electrodes in the arc chamber by a recombination of protons with atoms and molecules on the walls. It has been found empirically that glass or quartz walls inhibit the formation of the heavier ions.

Another closely related trouble encountered with this simple ion source is that of ion velocity distribution. For precision nuclear work it is important that the velocity distribution of the particles in the ion beam be much less than the desired error of the experiment. For example, at 2 Mev a voltage accuracy of 0.1 per cent is  $\pm 2,000$  volts. The arc potential in the ion source is less than 100 volts, so the possible voltage distribu-

tion of the ions coming from the arc chamber could not be greater than this value. Consider the case when the probe potential is high—say 10 kv. The probe canal is small and a region of high gas pressure; an ideal condition for the various charged particles to change their mass numbers. Protons going unchanged through the probe system from the arc chamber will possess the 10 kv of energy from the probe potential when they enter the ion acceleration tube. A mass 2 hydrogen molecular ion will also be accelerated by the probe gap to 10 kv, but if it should suffer a collision and break up in the probe canal into an atom and a proton, the 10-kv energy gained would be divided between the particles so that the resulting proton would have only one-half the energy of a proton passing unchanged through the probe canal. Other and more complicated processes may occur, resulting in a possible spread in the energy of the ions from the ion source and consequently at the target. This can be avoided in several ways; the probe voltage can be kept lower than the desired energy spread in the beam, or the high pressure region in the probe canal can be eliminated by ion source design.

The arc discharge of this simple type of ion source is, because of the low pressure, a very thin glow which occupies all available volume in the arc chamber, and is probably independent of the position of the filament and anode. At these pressures the mean-free path of the electrons from the filament is so long that those which produce the most protons are the ones which take the longest

path. This tends to make the density of the discharge very low. An attempt to concentrate this discharge by restricting the area through which it can travel is made in the Zinn-type<sup>7</sup> ion source, Fig. 5. The discharge between the filament and the anode is restricted by an anode shield, which is at the same potential as the filament. This shield forces the discharge into a small volume in a region near the arc chamber outlet so that more ions can be extracted by the probe. This type of ion source has been fairly satisfactory but it requires a large gas flow and high arc current to keep it operating. The engineering problems resulting from the heat developed by this large arc current make the source difficult to use in a pressure-type generator.

Another type of ion source whose arc chamber design is based on the same principles as the Zinn source is that developed at the Massachusetts Institute of Technology by W. W. Buechner, E. S. Lamar, and R. J. Van de Graaff.<sup>8</sup>

One of the many modifications of this type of ion source is shown in Fig. 6. The arc chamber design is an extension of the Zinn source in that the arc formed between the filament and anode must travel through a small restricted capillary instead of a cylindrical slit. The arc discharge will therefore be much more concentrated at the arc chamber exit hole. In the design shown in Fig. 6, however, no true probe has been used and the ions are allowed to drift out of the discharge chamber by themselves into a differentially pumped low pressure region. This eliminates the high pressure region in the probe canal with its particle mass-number change and the resulting wider spread in ion velocities from the ion source. The ions,

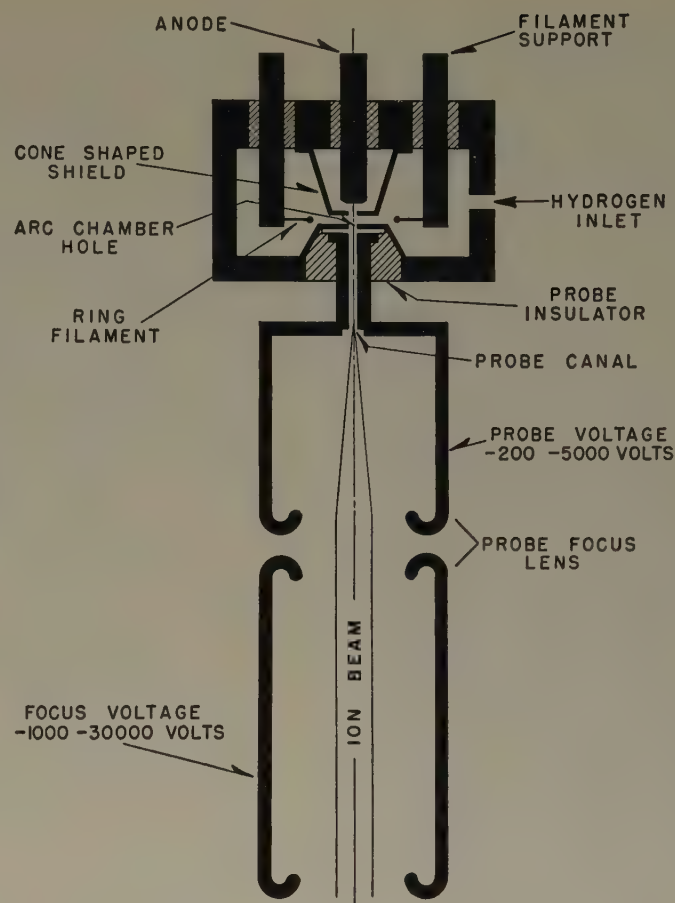


Fig. 5—A Zinn-type ion source.

drifting out of the arc chamber in a very divergent cone, are focused by the field of the first and second cylindrical focusing lenses through the exit hole into the main accelerating tube. The exit hole is small enough

to restrict gas flow into the main tube so the differential pumping system may remove most of the uncharged gas molecules which leak out of the discharge chamber. This has been modified in several ways, one

<sup>7</sup> W. H. Zinn, "Low voltage positive ion source," *Phys. Rev.*, vol. 52, p. 655; September, 1937.

<sup>8</sup> W. W. Buechner, E. S. Lamar, and R. J. Van de Graaff, "Production of proton beams," *Jour. Appl. Phys.*, vol. 12, p. 141; February, 1941.

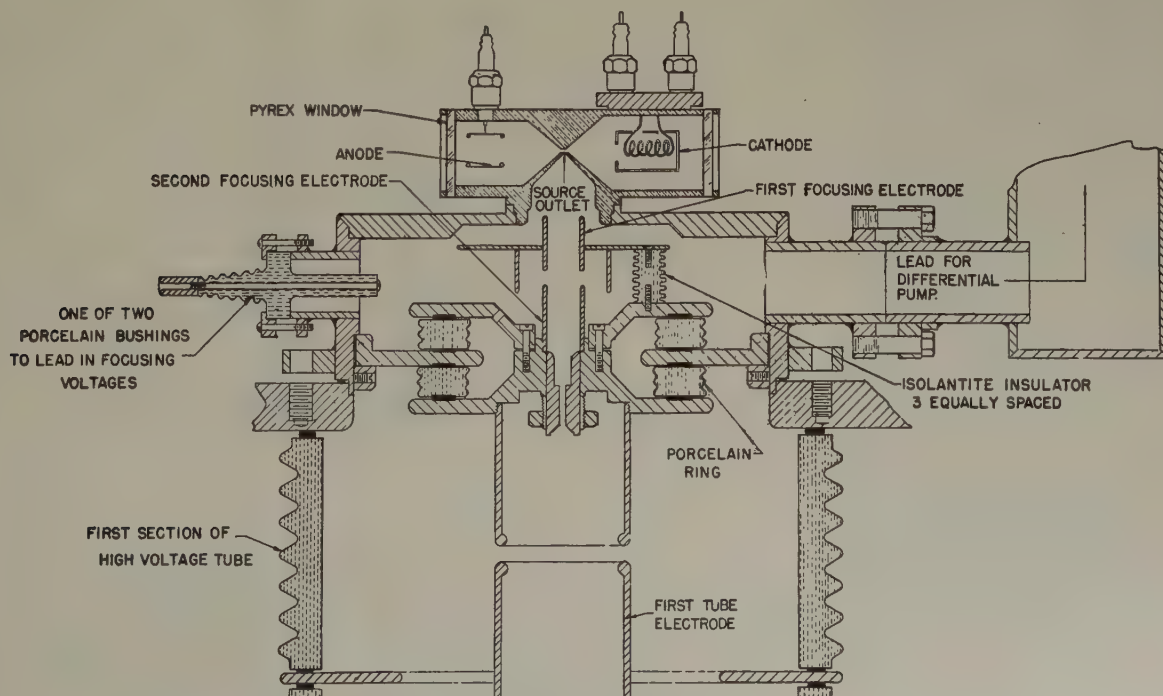


Fig. 6—A capillary-type ion source.

of which is to use a glass capillary for the discharge chamber. The percentage of protons over particles of other masses can be increased in this manner, although at the cost of some serious engineering difficulties in both the construction and use of the ion source. One modification of this source is to be used in the new MIT 12-million volt electrostatic generator.

#### RADIO-FREQUENCY ION SOURCES

Ion production by an electrodeless discharge<sup>9</sup> is interesting for two reasons. It eliminates the filament (see Fig. 7) and its attendant troubles, short life and difficulty in replacement. Such an electrodeless discharge can be operated readily in a glass or quartz discharge chamber which inhibits the recombination of protons to heavier molecular ions at the walls. The rf ion source described here was designed by R. N. Hall at the California Institute of Technology, and is powered by a 60-watt 400-Mc oscillator. The arc chamber is a small pyrex cylinder which is part of a tuned rf cavity of the coaxial type. An adjustable plunger in this cavity tunes it to the frequency of the oscillator. An axial magnetic field supplied by an external magnet coil effectively lengthens the paths of the electrons oscillating in the arc discharge chamber under the influence of the rf field by making the electrons take spirals paths. This effectively lowers the gas pressure at which the ion source will operate. The ions are allowed to drift out of the discharge chamber, and are focused by a probe potential in a manner similar to that used in the MIT capillary ion source. The ion source shown in Fig. 7 is a test source only, and requires further engineering before it could be used in a pressure generator.

#### ION SOURCES CONCLUSION

The advantages of the various types of ion sources previously described are all in the direction of larger proton currents with less gas flow into the main accelerating tube. The development of the differential pumping system is by far the most outstanding improvement in the last few years.

Many ion sources have one characteristic in common—they operate amazingly well on an ion source test bench with a short accelerating tube and very poorly when installed in an electrostatic generator with a long acceleration tube. Numerous papers have been published on test bench performances of ion sources but very few have discussed the application of the same ion source to an electrostatic generator. It would seem self-evident that if a beam starts out in the direction of a distant target and is well focused, it should all get to the target. This does not appear to be borne out in practice. The proton current from most generators is very low (of the order of one  $\mu\text{a}$ ) while the ion sources have been reported to give 20 to 1,000  $\mu\text{a}$ . One empirical fact has been noticed; generators with short acceleration tubes like the "short tank" generator at Los Alamos can produce very sizeable ion beams of the order of 50  $\mu\text{a}$  or more. The very long

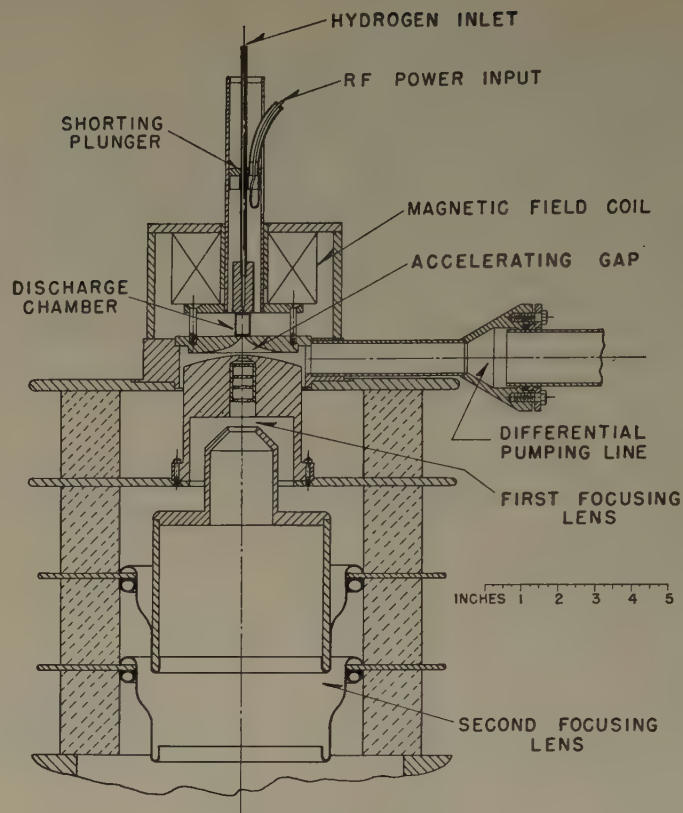


Fig. 7—Rf ion source.

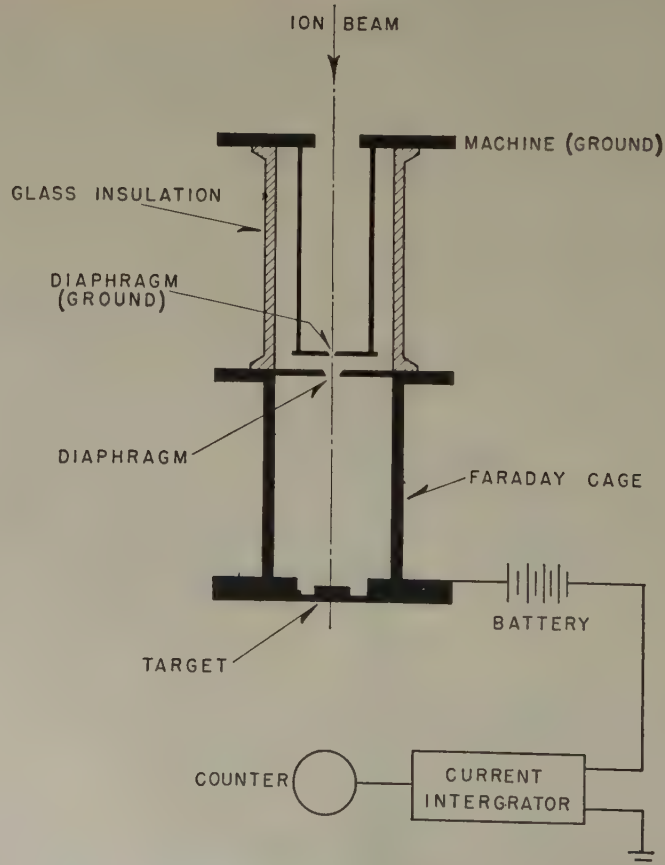


Fig. 8—A beam current collection and measurement.

<sup>9</sup> Robert N. Hall, "High frequency proton source," *Rev. Sci. Instr.*, vol. 19, p. 905; December, 1948.

tube machines like the generators at the Westinghouse Research Laboratories and the Carnegie Institute of Terrestrial Magnetism operate generally at levels of 1  $\mu$ a. The ion current carried by the tube seems to be limited by the tube length, but it is not easy to understand why. The fact that a longer tube has a slow pumping speed and hence less favorable vacuum conditions does not completely explain this phenomena. Further work should be done on this part of the design problem.

ION CURRENT MEASUREMENT

The ion beam originating in the ion source is accelerated through the vacuum tube by the potential developed in the generator. The beam, consisting of protons, and atomic and molecular ions must be analyzed into its various mass components so that only a homogeneous proton beam is retained. This is done by a magnetic field so that ions with the least momentum are bent through the greatest angle and sent into the target chamber. One further problem remains in this field, and that is to measure the ion current striking the target.

It is desirable in most nuclear experiments to know the total number of ions striking the target during a length of time as long as many hundreds of seconds. For this reason an electronic current integration circuit is desirable, instead of a microammeter or galvanometer. The current integrator is usually nothing more than a condenser, charged by the beam current. As current flows into the condenser, the voltage across it rises until at some predetermined fixed voltage the condenser is discharged, and the cycle starts over again. A simple gas triode can be used to discharge the condenser or, in some of the later type circuits, it is done with a multivibrator system. It is only necessary to count the number of times the condenser is charged during the operating interval to know how much current has struck the target. This is done simply by counting the condenser discharges with some type of an electronic circuit like a scaler.

The proton current from the generator is usually collected in an enclosed Faraday cage, shown in Fig. 8. A Faraday cage is a totally enclosed hollow conductor, so that all charges appearing on the inside run to the outside surface. It is not a bad approximation of a true Faraday cage to have a hole small compared to its dimensions through which the focused ion beam may enter. The charges carried by the ion beam, therefore, flow to the surface where they can be removed by a wire connected to the current integrating circuit. A second and very important reason for the use of such a cage is that the ions produce many secondary electrons when they strike a target. If these electrons could get out of the cage, their charge would be measured by the current integrating equipment as if it were from protons coming in. As there may be many electrons produced per single proton, large errors can be made in this way. It has been found, in fact, that even a fairly long Faraday cage is not enough to eliminate this error, so a small electric field is established between the cage and ground to push the electrons back into the cage where they were formed. A potential of about 300 volts is

usually enough for most applications, but where high precision is necessary an additional magnetic field across the cage can be used to deflect the electrons into the walls of the chamber.

GENERATOR VOLTAGE MEASUREMENT

It has been shown that the measurement of the voltage at which some nuclear process, such as a threshold or resonance level, occurs is of greatest importance. In practice, such a measurement is made by counting the number of particles resulting from the target bombardment for a considerable period of time at one voltage and repeating this for many other voltages to obtain a curve of yield versus energy. This experimental procedure places two requirements on the design of the voltage measuring and control equipment. It must be able to hold the generator operating voltage to the required value for long periods of time and permit this voltage to be easily and quickly changed. For these reasons voltage stabilizing feedback circuits are used, with the voltage measuring equipment as the source of error signal for some type of voltage control mechanism.

This routine voltage measuring equipment is only used for *relative* voltage measurements, comparing the voltage of the nuclear process in question with some secondary standard and assuming linearity of the measuring equipment.

The equipment used for the *absolute* voltage measurement of these secondary standards will be described first. The calibrated electrostatic analyzer and the radio-frequency ion velocity gauge are in this class. The uncalibrated electrostatic analyzer and the generating voltmeter are used for *relative* voltage measurement and error signal sources for stability control. Finally, the methods of generator voltage control from these error signals will be described.

ABSOLUTE VOLTAGE MEASUREMENTS

No method has as yet been devised for directly measuring electrical potentials of the order of millions of volts with an accuracy of better than about 10 per cent. We have, however, simple instruments which are capable of making *relative* measurements of such high potentials to a much better accuracy so that the practice has been to select a few secondary voltage standards and interpolate voltages from these. Such a secondary voltage standard must be easy to set up in any laboratory having an electrostatic generator and be easily reproducible. Nuclear physicists have turned to well-known nuclear processes for such a standard, and the most important one used today is the threshold of the production of neutrons when lithium is bombarded with protons which occurs at 1.882  $\pm$  (0.002) Mev. This reaction is very sharp, experimentally easy to do, and can be used in any laboratory for voltage calibration. There are several other voltage standards, such as the Al(*p*,  $\gamma$ ) resonance at 0.9933 Mev, and the F(*p*,  $\gamma$ ) resonance at 0.8735 Mev, but these are not as widely used. At present the only absolute voltage measurements that have been made were on these secondary standards. All nuclear physics measurements are made relative to one or the other of these calibrated points. The only methods of absolute volt-

age calibration available at present are based on the properties of the ion beam after it has been accelerated by the generator potential. Two methods have been applied to the measurement of the Li(*p*, *n*) threshold: that by deflection of the ion beam in an electrostatic field and by measurement of the velocity of the beam after acceleration by radio-frequency means.

It is interesting to point out that although electrostatic generators throughout the country which are used to accelerate protons have an accurate voltage scale, the electron generators do not. So far there is no direct correlation between the voltage scales used by the two types of generators, except a remote one through various nuclear processes of questionable accuracy. This correlation between the two scales is very important and merits further investigation.

The ion beam, being made up of charged particles with a well-known *e/m* ratio, can be deflected by electric or magnetic fields or both. If the field strengths are accurately known, the radius of curvature of the path of the ions of any given velocity or voltage may be calculated. Slit systems can be used to establish any desired radius of curvature so that the ion beam voltage can be determined if the field strength is known. At present the production of magnetic fields, constant and known over a large area, to better than 0.1 per cent is so difficult that this method has not been used. However, many of the necessary techniques have been developed by the designers of cyclotron magnets so that magnetic analyzers may soon take their place in the precision voltage measurement field. Electrostatic analyzers, however, have been used with high accuracy. The electric and magnetic analyzers have a great advantage in that by using slit systems they can select only ions of a certain voltage having a known voltage distribution.

A particle of charge *e* and mass *m* has an energy *E* and a velocity *v* after being accelerated by a potential *V* given by the equation:

$$E = 1/2mv^2 = eV. \tag{1}$$

In precise calculations a correction term must be added to this equation for the relativistic change in mass with velocity, but we may neglect it in this simple description of an electrostatic analyzer. If a small parallel beam of particles of energy *E* is passed between two plates, Fig. 9, of length *L* and separation *d* and at a potential difference *v*, it will be deflected an amount *S*. The field  $\epsilon$  between the plates will be:

$$\epsilon = \frac{v}{d}. \tag{2}$$

And the force on the particles in the beam due to this field:

$$f = \frac{ev}{d} = ma, \tag{3}$$

where *a* is the acceleration. It will be accelerated for a time *t* given by:

$$t = \frac{L}{v}. \tag{4}$$

The deflection *S* at the end of the analyzer plates due to this acceleration will be:

$$S = 1/2at^2 = \frac{1}{4} \frac{e}{d} \frac{L^2v}{E}.$$

The deflection  $S$  is proportional to the inverse of the energy of the particle and by placing a slit of width  $\Delta S$  in the correct position only particles of an energy distribution  $\Delta E$  can come through and fall on the target.

It is difficult with a parallel plate analyzer to get sufficient energy resolution to make the instrument practical because the space  $d$  must be large enough to allow for the curvature of the beam. The field, Fig. 10, between plates curved to form a section of a cylinder of radius  $R_1$  and  $R_2$  will bend a beam of ions of constant energy in a circle of radius  $R$  between  $R_1$  and  $R_2$ . Much higher resolution may be obtained from this type of an analyzer because of the greater deflection angle. The radius of the ion beam may be established by a slit system at each end of the analyzer plates and the analyzer voltage adjusted to allow ions of a certain energy to go through.

This type of electrostatic analyzer,<sup>10,11</sup> was used by Herb, Snowdon, and Sala in the precision voltage measurement of the nuclear reactions generally used as secondary voltage standards. The plates of the analyzer are about 1.5 meters long and separated by about 1 cm. These plates were very carefully made the variations in plate separation are not over 0.0005 inch.

The potential applied to the analyzer plates must be known to a very high precision. The voltage required for the measurement of the  $\text{Li}(p,n)$  threshold was about 30 kv with one-half of this potential on each plate and was provided by two 15 kv stacks of dry batteries composed of several fixed battery boxes with an adjustable box at the low potential end. The voltage of one of the battery boxes was measured by precision dc methods, (voltage divider and type  $K$  potentiometer and standard cell) and adjusted to precisely 500 volts. This box, now a secondary voltage standard, can be compared with another 500-volt box by adjusting the voltage of the second box until no circulating currents were found when they were connected together. The two known 500-volt boxes were added together and a 1,000-volt box set for zero circulating current against the sum of two 500-volt boxes. This was repeated until both stacks were adjusted to exactly 15 kv. Small voltage variations were made in one of the 500-volt boxes and measured in the usual dc manner. It was felt that the total voltages of these battery stacks were known to an accuracy of 0.03 per cent.

It is interesting to note that electronic controlled power supplies were not used because of voltage instability. However, an electronic power supply controlled by a galvanometer and photocell system has proved to be more stable than the battery boxes.

Knowing the analyzer potential, the radius of curvature of the beam path as established by the slit system, the plate separation, and making the necessary corrections for relativity and other minor second-order effects the voltage of the generator may be

<sup>10</sup> R. G. Herb, S. C. Snowdon, and O. Sala, "Absolute voltage determination of three nuclear reactions," *Phys. Rev.*, vol. 75, p. 246; January, 1949.

<sup>11</sup> R. E. Warren, J. L. Powell, and R. G. Herb, "Electrostatic analyzer for selection of homogeneous on beam," *Rev. Sci. Instr.*, vol. 18, p. 559; August, 1947.

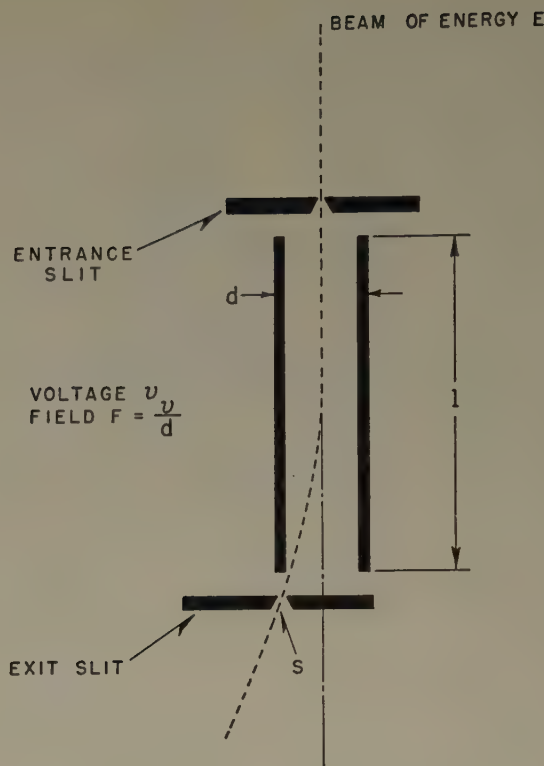


Fig. 9—A parallel plate electrostatic analyzer.

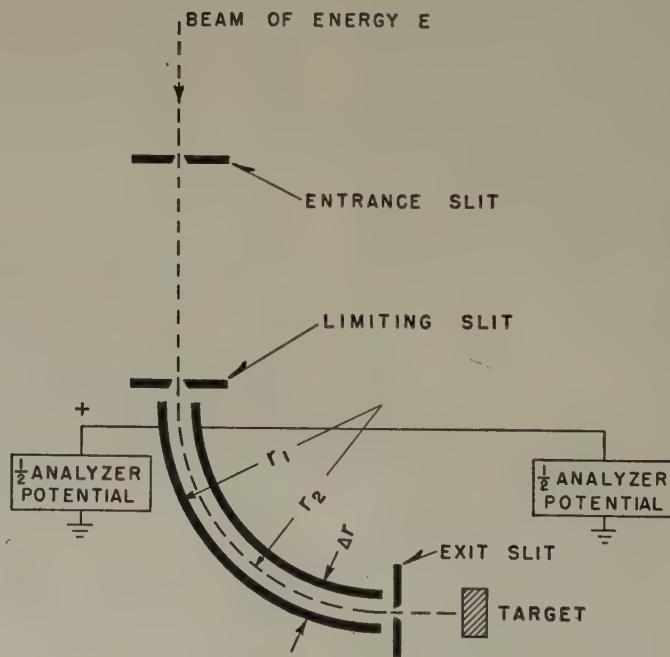


Fig. 10—A curved plate electrostatic analyzer.

obtained. As the analyzer plate voltage is now a measure of the voltage of the ion beam which will come through the analyzer system, it may be used as a primary voltage scale. The measurement of a secondary voltage standard such as the  $\text{Li}^7(p,n)$  threshold involves the recording of the number of neutrons produced by the target as a function of beam voltage at various voltages close to the threshold value. The threshold can be

determined by selecting the highest voltage at which no neutrons appear. This was found to be  $1.8820 \pm 0.002$  Mev.

#### ABSOLUTE VOLTAGE MEASUREMENT BY THE RADIO-FREQUENCY ION VELOCITY GAUGE

This method, which also makes use of the known constants of a beam of ions, permits

one to establish a number of voltage calibration points along the nuclear voltage scale.<sup>12,13</sup> It is absolute in the sense that the kinetic energy of the particle is determined in terms of its known mass and charge and the directly and accurately measurable parameters—length and frequency. The ion beam from the electrostatic generator is intensity modulated by a beam defocusing electrode placed just below the ion source and supplied by a 70-Mc crystal-controlled oscillator mounted in the high voltage electrode. This beam of modulated ions is accelerated by the generator potential and analyzed into its mass components by a momentum analyzer. The desired beam is then sent through two gaps in an rf cavity as shown in Fig. 11. This cavity (Fig. 12) is of the coaxial type, with the inside of the center conductor used as a field free drift tube. The cavity can be tuned to resonate at the beam modulation frequency by the gap condensers. The ion beam, as it passes through the first gap in the rf cavity, induces a 70-Mc voltage in the cavity because of its intensity modulation. As the rf cavity has a high  $Q$ , the two gaps are closely coupled and an rf measuring instrument will indicate the sum or difference voltage of the two gaps. The addition or subtraction of the voltage from these two gaps will depend on the phase difference of the 70-Mc modulation cycle between the two gaps. If the transit angle  $\Delta\theta$  is equal to  $n\pi$  (where  $n$  is an odd integer) the excitation from the first gap will be cancelled by that from the second and a minimum in rf signal will result. Whence

$$\Delta\theta = n\pi = 2\pi f\Delta t \quad (5)$$

$$\Delta t = n/2f = L/v, \quad (6)$$

where  $\Delta t$  is the transit time of any group of ions of velocity  $v$  through length  $L$  between the gaps, and  $f$  is the modulation frequency. The velocity of the ions will therefore be

$$v = \frac{2fL}{n} \quad (7)$$

and the voltage  $V$  of a beam of particles of velocity  $v$  and mass to charge ratio  $M_0/e$  will be

$$V = \frac{2M_0 f^2 L^2}{en^2}. \quad (8)$$

In practice it is necessary to include the relativity correction for the change in mass of the ion with velocity so that the final result becomes:

$$V = \frac{2Mf^2L^2}{en^2} \left( 1 + \frac{3f^2L^2}{n^2C^2} \right)$$

where  $C$  is the velocity of light.

With a fixed modulation frequency  $f$  and distance  $L$  between the gaps, it is possible to find several minima along the nuclear voltage scale with a precision of about 0.1 per cent by using various values of the order number  $n$ .

To make the measurement of the rf signal from the cavity easier, the 70-Mc modu-

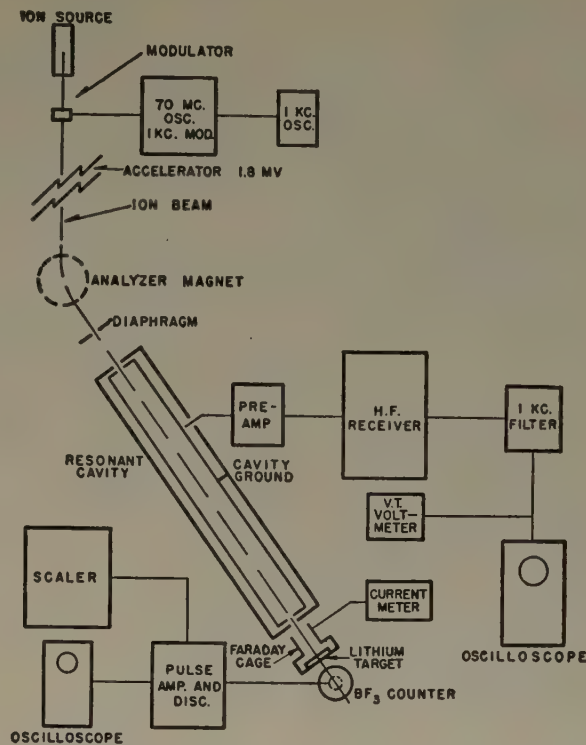


Fig. 11—Rf ion speed gauge assembly. The magnetic analyzer permits rapid switching from the proton beam used for threshold measurement to the mass 2 beam used for the speed measurement, the mechanical geometry remaining fixed.

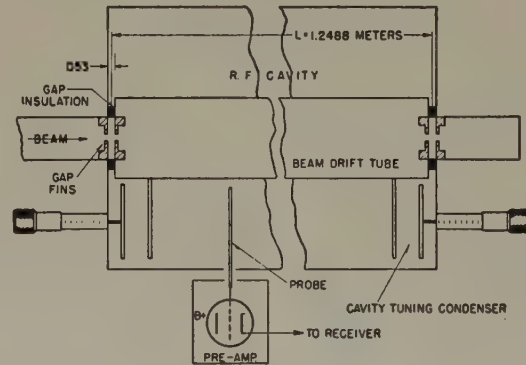


Fig. 12—A schematic section of the speed gauge cavity, giving essential dimensions.

lation voltage in the electrode is further modulated by a 1,000-cycle signal to allow audio amplifiers to be used after the 70-Mc detection circuits. The signal from the cavity is picked up by a voltage probe and fed to a preamplifier and impedance matching tube adjacent to the probe. It is amplified and detected by a communications receiver with a 1,000-cycle narrow-band filter in its output, which reduces the receiver bandwidth to the order of 100 cycles thus considerably increasing the signal-to-noise ratio. The output signal can be read on a scope or a meter. The magnitude of the rf signal from the cavity is recorded as a function of an arbitrary voltage scale as determined by a compensating generating voltmeter. A measurement of

the lithium<sup>14</sup> threshold is made on the same arbitrary scale. Knowing the constants of the cavity, the length between the gaps, and the modulation frequency it is possible to calculate the absolute voltage at which the minimum in the rf signal occurs. From the experimental points recorded a minimum in the rf signal can be selected which calibrates the arbitrary voltage scale in absolute volts. A linear extrapolation of this point can be made to the  $\text{Li}(pn)$  threshold. The value found in this manner was  $1.8812 \pm 0.002$  Mev which may be compared with  $1.8820 \pm 0.0002$

<sup>12</sup> W. Altar and M. Garbuny, and J. W. Coltman, "A speed gauge for high voltage ion beams," *Phys. Rev.*, vol. 72, p. 528; September, 1947.

<sup>13</sup> W. Altar, and M. Garbuny, "Absolute speed gauge for high voltage particles," *Phys. Rev.*, vol. 76, p. 496; August, 1949.

<sup>14</sup> W. E. Shoupp, B. Jennings, and W. Jones, "An absolute calibration of the  $\text{Li}^7(p, n)$  threshold voltage," *Phys. Rev.*, vol. 76, p. 502, August, 1949.

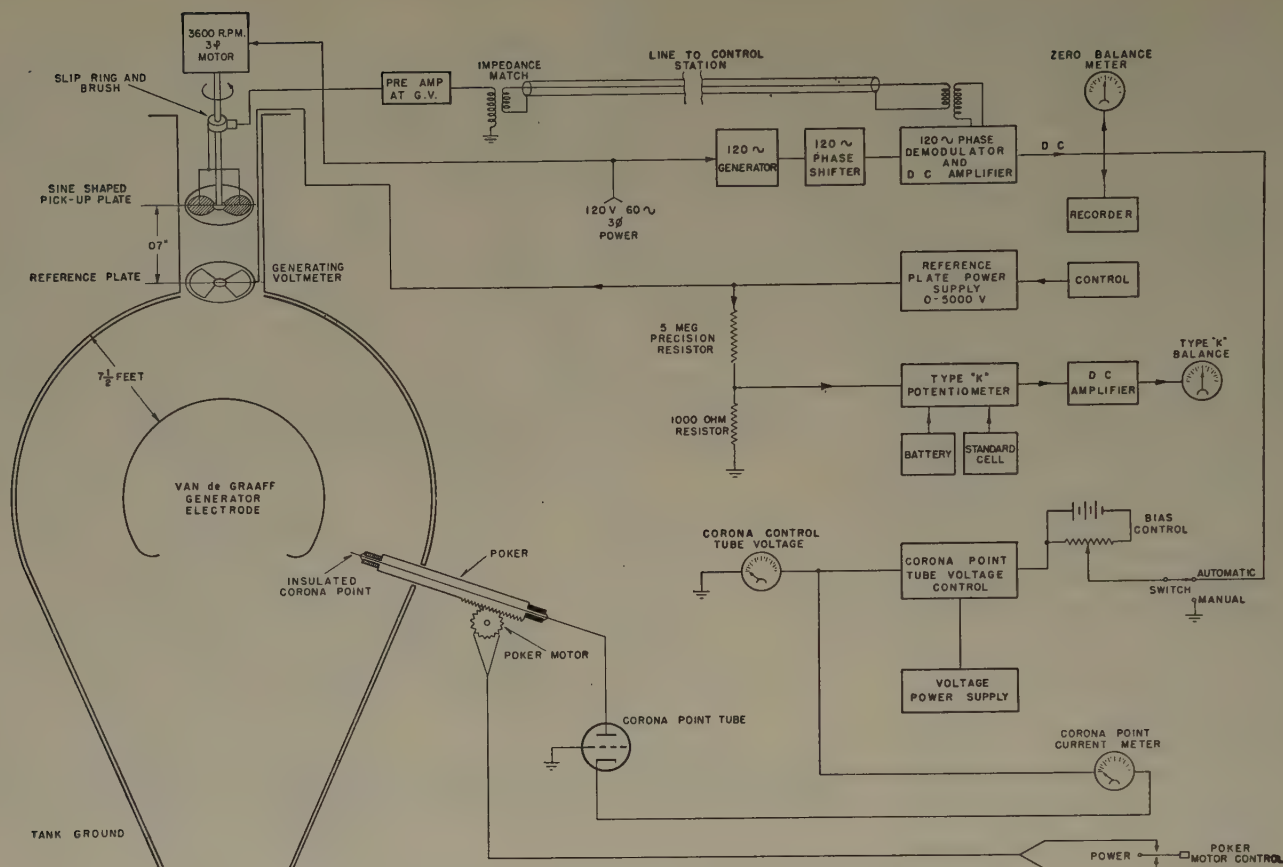


Fig. 13—A Westinghouse Van de Graaff generator generating voltmeter comparison and control circuits.

Mev, as determined with an electrostatic analyzer.

#### GENERATING VOLTMETER

The previously described instruments can be used for absolute calibration of the nuclear voltage scale. The generating voltmeter can be used for relative measurements only. This instrument is a field measuring device which determines the field strength due to the high potential electrode at the surface of the tank. Several types of generating voltmeters have been used, the simplest form consisting of an insulated segmented plate exposed to the field of the electrode. A mechanical rotating shutter at ground potential alternately exposes this plate to the field of the electrode and to ground so that a resistor between this insulated plate and ground has a small ac potential developed across it, whose peak value is a measure of the electrode field strength. This peak ac voltage as read by some form of vacuum-tube voltmeter therefore becomes a measure of the electrode voltage. However, this instrument is not capable of extremely high accuracy because of variation in generated wave shape.

Since a direct reading generator voltmeter involves the problem of accurately measuring an ac voltage of very poor wave shape, a null-type instrument would be advantageous. This is done in the compensat-

ing-type voltmeter which is capable of somewhat better precision than the direct reading instrument. As all these instruments (see Fig. 13) are of the electrostatic ac generator type, it must also have a rotating element or armature. This rotating element, in the more advanced design voltmeters, consists of a pair of sine-wave shaped insulated pick-up plates connected through a slip ring and a low impedance matching transformer to ground. An insulated stationary plate cut so as to leave open two 90° sectors is placed a short distance in front of the pickup plate. The whole is placed in the electrostatic generator so that it is exposed to the high potential field from the electrode. The rotating pickup plate continuously explores the field distribution behind the reference plate, being alternately exposed to the electric field from the electrode and to that of the insulated reference plate. If the insulated reference plate is charged with a dc potential of the same sign as the electrode, there will be some potential at which it will be on an equipotential plane and the field intensity from it will be the same as that from the electrode. When these fields are balanced, the rotating pickup plate will be in a homogeneous field and the ac generated will be, to a first approximation, zero. This null can be used to indicate that the dc reference plate potential is some unknown but constant fraction of the electrode voltage. Although the absolute value of this ratio is dependent on the rela-

tive spacing of the various components, the geometry is so complex that the ratio cannot be theoretically determined with any accuracy. However, it has been experimentally shown that the relationship is linear over very wide ranges of voltages to an accuracy of about a tenth of one per cent.

The signal generated by the pickup plate when the reference plate voltage and the electrode voltage are not in the correct ratio is an ac of twice the frequency of the mechanical driving motor and there is a 180° shift of this ac signal at the balance or null point. This is fortunate, as it allows the generating voltmeter to be used as a source of error signal for voltage stabilization circuits. For voltage measurement only, it is enough to present the output of the pickup plate on a high gain scope and adjust the voltage for minimum ac signal. The dc reference plate voltage can be measured to any degree of accuracy necessary by standard low-voltage dc measurement techniques, which is one of the major improvements in this type of voltmeter over the ac type. The limit of accuracy of this type of instrument is, of course, in the second-order effects. In the first place, the field from the electrode does not completely match that of the reference plate, due perhaps to edge effects of the plate itself and to mechanical inaccuracy in the equipment. This "noise" is of higher harmonics than the true ac generated, but it obscures the voltage minimum. This is not a limita-

tion, as one can always use the same wave pattern as a balance criteria, either by observation on a scope or by electronic means. However, it has been found that the wave shape of this noise does not remain constant, and this fact definitely limits the accuracy of the instrument.

#### METHODS OF GENERATOR VOLTAGE CONTROL AND STABILIZATION

The voltage of an electrostatic generator, even with the belt speed, charging current, and other input conditions held as constant as is practical, is still subject to considerable variation. There is a fast voltage change probably associated with the surface conditions of the charging belts and with the local corona conditions on the electrode and down the support members. There is also usually a slow drift in voltage which is due to an integrated change in all these conditions. The fast voltage variation may not amount to more than a few tenths of one per cent, but the slow drift may be as great as one per cent or more during a long period of operation.

The voltage at which an electrostatic generator operates is governed by current taken up the charging belt and that lost by all paths down to ground. To control the generator voltage, one must be able to regulate one or both of these currents. It has been found that controlling the electrostatic generator voltage by means of the belt spray current is not satisfactory except as a slow control, because there is considerable time delay in conveying the charge to the electrode on the relatively slow belt.

If a sharp point or group of points is inserted through the ground wall of the electrostatic generator tank (Fig. 14) it will become a source of corona current to the electrode. The corona current can be controlled either by mechanically adjusting the distance between the point and the electrode, or by changing the potential of the corona point with respect to the tank wall. Forcing this corona point to go several tens of kilovolts above ground has the effect of redistributing the field lines from the electrode so that the field strength at the discharge point is very much less. If the corona point is insulated from ground and connected to the plate of a high-voltage triode vacuum tube whose cathode is grounded, the grid of this tube can be biased to control the current to the corona point, and hence its voltage. In this simple manner a fine, relatively fast control of the current lost from the electrode, and hence the electrode voltage may be obtained.

Another method of controlling the loss of current from the electrode is by the process of shooting a stream of electrons (Fig. 15) from ground to the electrode and controlling the magnitude of the electron current. This method, although very fast, usually requires a second vacuum tube to be installed in the generator for the electron stream. Also the very high voltage X rays produced when these electrons are stopped at the top of the tube can be hazardous from a health standpoint.

A third method which has been used on the electron generator at Notre Dame is that of capacitance control. The inside of the electrostatic generator tank (see Fig. 1)

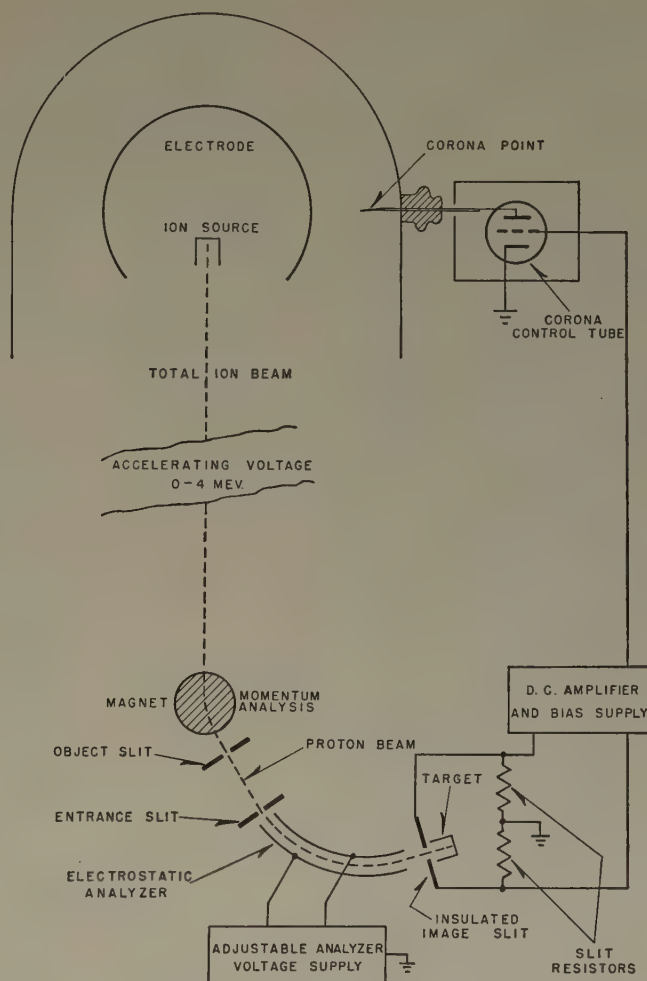


Fig. 14—A corona point voltage control using electrostatic analyzer.

has a metal lining insulated from ground for about 100 kv. The potential of the generator electrode is now with respect to the lining and not ground, although the ion beam is accelerated by the potential from the electrode to ground. If the potential of the tank lining is raised above ground, the electrode potential with respect to ground will be increased. This method is limited by the fact that it cannot compensate for long slow drifts in machine voltage, but these can be taken care of by one of the other slower control methods. Its response to a voltage error signal is fast, being limited only by the power of the liner voltage control amplifiers in changing the voltage of the liner. This method has also been proposed for the new 12-Mev generator at Los Alamos.

#### ERROR SOURCE FOR VOLTAGE STABILIZATION

If the two exit slits described in the section on electrostatic analyzers and shown in Fig. 14 are insulated from ground and from each other, the potential developed across a resistor from each slit to ground will be a measure of the ion current striking it. If the ion beam velocity or voltage changes a small amount from that for which the electrostatic analyzer voltage is set, more of the

ion beam will strike one slit than the other and a difference in the slit resistor voltage drops will appear. These voltage differences can be amplified by dc electronic amplifiers and used to bias the grid control tubes in either of the previously described voltage stabilizing circuits. This forms a feedback loop or servo system in which analyzer voltage is the standard and slit resistor voltages are the error signals operating the generator voltage control mechanism. This mechanism, by changing the generator voltage in the direction to reduce the error signal, completes the loop. To change the operating voltage of the electrostatic generator, at least within the rather narrow limits of the control mechanism, it is only necessary to change the electrostatic analyzer plate voltage and the servo system will cause the generator voltage to follow.

#### GENERATING VOLTMETER VOLTAGE STABILIZATION METHODS

The compensated generating voltmeter as described in the section of voltage measurement (Fig. 13), produces a minimum in the generated ac error signal when the reference plate potential is on an equipotential plane of the electrode. When the voltage of the electrode changes, a 120-cycle ac is produced

whose phase shifts  $180^\circ$ , dependent on whether the electrode voltage is higher or lower than the balance voltage. This phase shift can be made use of in voltage stabilization of the electrostatic generator.

While it would be very simple to convert this ac voltage to dc by a simple rotating commutator as it is done in a two-pole dc machine, it is better to amplify it to a much higher level while it is still capable of being handled by ac electronic amplifiers. In this amplification process the signal can be cleaned up by narrow-band 120-cycle filters while it is still in the ac form. Although these filters introduce some signal phase shift of their own, this is not serious when one remembers that the rate of change of generator voltage is limited (except at spark over) by the fact that the current drawn from the electrode by any control system is very small and the time to reduce the generator voltage to  $1/e$  of its former value may be as great as 10 seconds.

The amplified ac null signal can then be demodulated against a reference ac whose phase is known and controlled from the rotation of the pickup plate. This ac can come either from a small ac generator rotating on the same shaft as the pickup plate or from the ac power driving the synchronous motor. An ordinary demodulator or synchronous amplifier can be used to derive a dc voltage whose sign is either positive or negative, depending on the relative voltages of the electrode and the reference plate. Amplification of this dc and additions of a constant bias voltage can be used to control the grid of a heavy transmitting tube in series with the corona point voltage control. At the same time the output dc from the demodulator serves as an indicator to the operator that the average voltage of the electrode has not changed and that his reference plate potential is still a measure of the electrode voltage. It must be emphasized that all the voltage measurements made in this manner are *not* absolute, and the instrument must be calibrated in some manner.

#### MISCELLANEOUS USES OF ELECTRONICS

There are a few minor uses of electronic circuits in connection with an electrostatic generator, such as ionization gauges, for measuring the ion tube vacuum. As the vacuum conditions found in an electrostatic generator ion tube are not extreme, conventional circuits can be used with the ionization gauge and are not of special interest. Relay systems are also used with the output of the ion gauge, so that if the vacuum does go bad for some reason, the gauges and pumps will be automatically turned off. These safety systems, although of great importance to the operation of the machine, are

again as simple and direct as they can be made to reduce accidental troubles and maintenance.

Some electrostatic generators have been designed with complicated telemetering systems involving either radio or light links between the electrode and ground for making measurements of the various voltages associated with the ion source. It is doubtful if

microamperes of protons with energies up to about 4 Mev. With great care, voltage accuracies of about 0.1 per cent are available. Even within this limited range, this generator can and has been used for many important nuclear investigations. The electrostatic generator is as close to the ideal design of generator for most nuclear experiments as can be conceived. For this reason its limita-

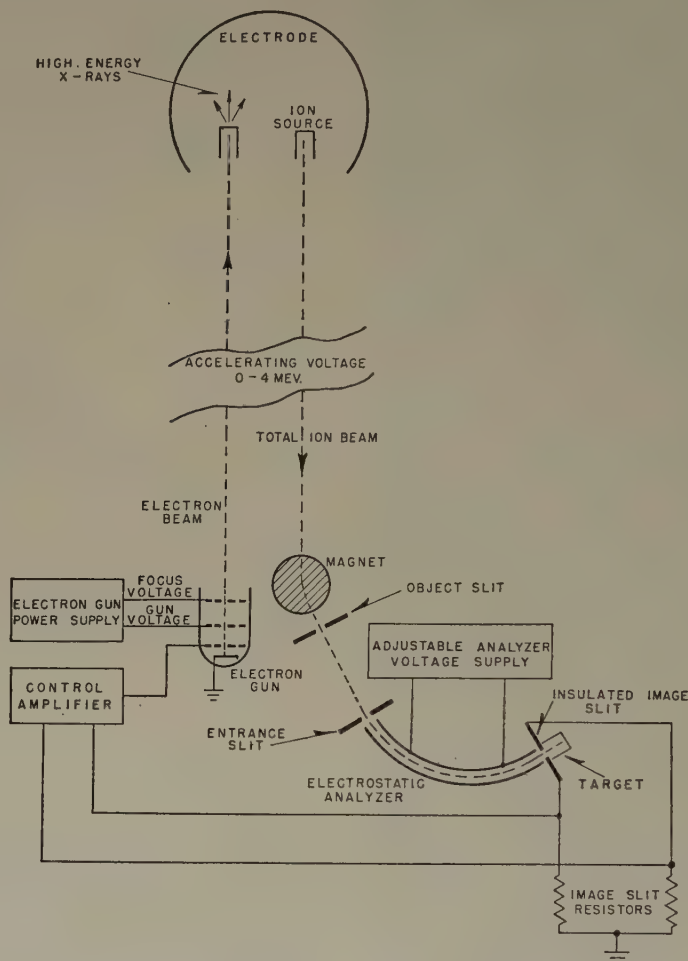


Fig. 15—A electron beam voltage control using electrostatic analyzer.

such systems can compete with the simple process of looking at meters in the electrode with a telescope.

#### CONCLUSIONS

The electrostatic generator, at the present stage of design, can produce a very few

tions, the small ion current, low top voltages, and poor voltage stability should be carefully studied.

The high voltage engineering groups at MIT and Los Alamos are engaged in just this task and their development of the proposed 12-Mev machines is of great interest.



# New Techniques for Electronic Miniaturization\*

ROBERT L. HENRY†, ASSOCIATE, IRE, ROBERT K-F SCAL†, MEMBER, IRE,  
AND GUSTAVE SHAPIRO†, ASSOCIATE, IRE

**Summary**—The general approach to the problems of miniaturization is outlined with special emphasis on problems caused by the high operating temperatures. In addition to the general techniques, both conventional and printed circuit, three models of miniature wide-band, high-gain radar-type intermediate-frequency amplifiers (having similar characteristics) are described. A discussion of production methods and techniques is also included.

## INTRODUCTION

THE RECENT phenomenal increase in the application of electronics in the fields of ordnance, aeronautics, and instrumentation has emphasized the need for reduction in size and weight of equipment. Furthermore, high-level performance under ambient conditions of increased severity has also been required.

To investigate the possibilities of electronic miniaturization, a program was instituted within the Engineering Electronics Section of the National Bureau of Standards sponsored by the Navy Bureau of Aeronautics with some contribution toward the later work by the Navy Bureau of Ordnance. This program had as its objective the development of new techniques and methods aimed toward miniaturization and toward development of mass-production methods. The embodiment of the results of some of these developments into certain specific electronic assemblies was also undertaken, and three of these assemblies are described. All are intermediate-frequency amplifiers of the same general type. The developments in this family of equipment illustrate some of the most promising techniques for miniaturizing electronic assemblies.

Early electronic miniaturization efforts were aimed only at utilizing standard components in as compact an assembly as possible. The method undertaken at the National Bureau of Standards has been to go farther and develop smaller components and standardized techniques which facilitate the production of component assemblies whose characteristics and circuitry may be conveniently varied for specific applications. The design of miniature equipment must satisfy not only the usual requirements associated more specifically with component design but also the conditions of practicality in terms of manufacturing, testing, and repairing difficulties. These problems require the attention and cooperative efforts of the chemist and mechanical engineer as well as the electronics engineer.

The question of how far miniaturization efforts should go is quite appropriate. The program of miniaturization of intermediate-frequency amplifiers at the National Bureau of Standards has demonstrated that the process can be carried to a point where the complete device is

only slightly larger than the volume occupied by the subminiature type tubes. However, such a degree of miniaturization is purchased at the price of greater difficulty in construction, and the miniaturization aspects of design must be adjusted to the requirements of the application if they are to be attained economically. There is, of course, the possibility that mass-production methods may make it economical to use the smallest components for all installations. This might be viewed as the ultimate goal of the "miniaturization" engineer.

A factor of great concern in the miniaturization of electronic equipment is the high internal temperature caused by packaging heat-dissipating components into small volumes. This problem is intensified by the recent tendency of military agencies to specify operation at much higher ambients than in the past. Some specifications already call for 100°C operation, and there is some indication of a desire to extend this to 200°C operation in the case of certain airborne equipment. These are formidable requirements. For example, conventional components are limited by their composition to operation in ambients of only 85°C or in some cases of 125°C. Yet measurements made within miniature intermediate-frequency amplifiers indicate that components sometimes operate in 200°C ambients when the amplifier is subjected to temperatures of only 85°C. The engineering of the miniature intermediate-frequency amplifiers has thus been greatly influenced by the desire to permit operation of the equipment in ambients approaching 85°C.

With these temperature problems in mind, the Model II amplifier, which was originally designed to use conventional resistors, was modified to permit utilization either of these or of special resistors for high-temperature operation. This latter resistor type was used exclusively in the Model V amplifier. It consists of a steatite rod on which a cracked carbon film is deposited and which is equipped with axial leads. For the printed circuit unit, a special high-temperature film-type resistor was developed in the National Bureau of Standards Laboratories.

High temperature also required the use of special solder having a melting point of about 220°C and the inclusion of sufficient silver to prevent alloying of fired-on silver from printed circuits to which connections were being soldered. This solder was used with the same general equipment and techniques, and with only slightly more difficulty than ordinary solder.

## CIRCUITRY

The selection of the high-frequency, wide-band intermediate-frequency amplifier as a vehicle for the evalua-

\* Decimal classification: R380×R363.4. Original manuscript received by the Institute, April 6, 1950.

† National Bureau of Standards, Washington, D. C.

tion of miniaturization techniques in the early stages of the program was predicated upon a number of considerations. Principal among these were: (1) an immediate need for such equipment by some government agencies, (2) the diversity of problems imposed by this type of equipment, and (3) the probability that the general solutions arrived at would have wide applicability in other types of equipment. The selection of specific circuitry for the intermediate-frequency amplifiers requires that all factors bearing on miniaturization, including the effect of high temperatures on components, must be considered and that prime emphasis must be placed on attainment of electrical requirements rather than extremes in miniaturization.

After consideration of various types of intermediate-frequency circuitry, it was decided that a stagger-tuned interstage coupling system using bifilar unity-coupled inductors was most adaptable to a miniature design. Such a system eliminates the need for coupling capacitors and provides excellent isolation of the various intermediate-frequency ground-return circuits. The use of staggered pairs rather than triples or quadruples provides less critical circuitry. Chain decoupling of power supplies was selected for Models II and PC-IV because of greater efficiency even though this requires the use of chokes rather than resistors in the decoupling filters. The above selections, together with the basic specifications for performance, defined the circuitry as shown in Fig. 1.

The pentode amplifiers are subminiature equivalents of the type 6AK5 miniature tube. Each of the models has been built with both the Sylvania type SN973B and the Raytheon type 5702 tubes. Although the former is a button-base type and the latter a press-base type, the physical layout of each design permits the use of either type without modification. The same tube types were used in the intermediate-frequency, video-amplifier, and cathode-follower stages; the duo-diode used in Models II and PC-IV is a National Union type 1106; and the single diode used in Model V is a Raytheon type 5704.

## MODEL II

The Model II intermediate-frequency amplifier was designed so that it could be produced by electronic equipment manufacturers without radical changes in their manufacturing methods. It was also designed to be sufficiently flexible so that modifications of circuitry and operating frequency could be made easily. Another objective was an extremely rugged and rigid design.

This amplifier has a 10-Mc bandwidth centered at 60 Mc. As shown in the schematic diagram, Fig. 1, it

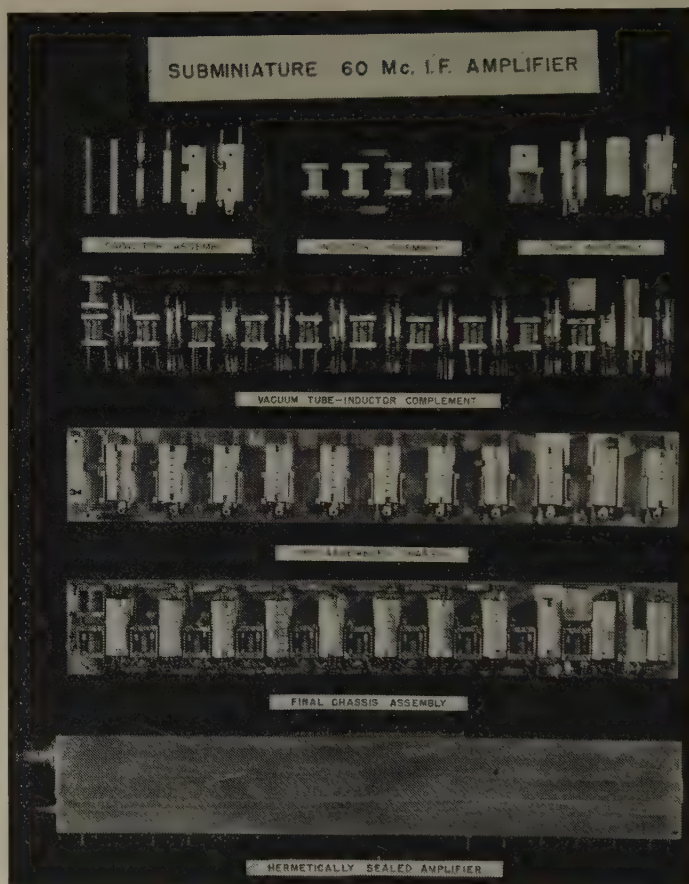


Fig. 2—Model II miniature subminiature amplifier.

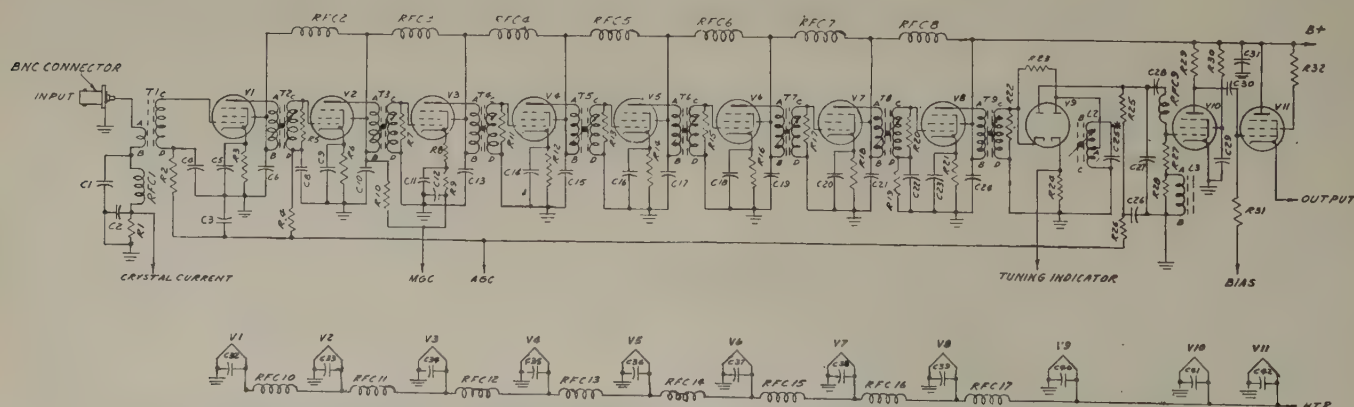


Fig. 1—Basic schematic diagram used to construct miniature wide-band amplifiers. Note—All transformers except  $T_1$  are bifilar wound.  $C_{25}$  is temperature compensating.

consists of an eight-stage stagger-tuned intermediate-frequency amplifier, diode detector, video amplifier, and cathode follower—eleven vacuum tubes in all. The gain-bandwidth factor approximates that of an amplifier using type 6AK5 tubes.

The amplifier is housed in a hermetically sealed plug-in container (see Fig. 2). Sealing is accomplished with the aid of a soldered-down metal strip that may be removed, in the event that the amplifier must be repaired.

The original intent was to pot this amplifier in a casting resin. However, there are some very serious objections to potting this unit. First, most of the existing casting resins cannot stand high temperatures. Second, the sealing provided by casting resins is not generally as satisfactory as that provided by a hermetically sealed container; there is a tendency for moisture creepage to occur along wires emerging from potted assemblies exposed to humid atmospheres for long periods. Third, dielectric constants of even the best casting resins being in the order of 2.5, the interstage capacity might be increased seriously and the over-all performance deteriorated by casting. In order to provide good metallic bonding to assist in heat transfer to the outer surface of the assembly and to provide low inductance grounds for many of the components, a mechanical structure was provided which proved to be sufficiently rugged that little additional rigidity resulted from the use of a casting resin.

Because miniature assemblies can require a great deal of skill on the part of the wiremen, interconnections made with hook-up wire were almost completely eliminated in the Model II design. Component leads and printed wiring were used for interconnections. The only other wires used were a shielded automatic-gain-control line and a 2-inch long B+ lead. Most of the wire connections were less than  $\frac{1}{4}$  inch long.

Intermediate-frequency amplifiers of this general type contain more by-pass capacitors than any other component. These capacitors must be located to permit efficient by-passing action for the intermediate-frequency portions of the circuit. The physical shape of the capacitor, therefore, plays a very important part in the determination of the intermediate-frequency layout. It is not necessary that a capacitor used for by-passing have a particularly good power factor as long as the total impedance at the operating frequency is low. In high-gain designs, low-loss by-pass capacitors often cause trouble by creating a condition conducive to parasitic oscillation. Since low- $Q$  by-pass capacitors are acceptable in this amplifier, it is possible to use some of the higher- $K$  barium titanates with high dissipation factors. Because the temperature characteristics of high- $K$  ceramic make it impossible to use resonant by-pass capacitors, high-capacity by-pass structures with low self-inductance were used. These took the form of  $\frac{1}{8}$ -inch-diameter thin-wall tubing silvered inside and outside.

The Model II chassis is a single metal plate to which the components are secured. The tubes lie parallel to each other and the coupling networks lie between the tubes. The tube mounting also functions as a tube and interstage shield and as a heat-transfer medium that bonds the vacuum tube to the outer wall of the shield case. The tube shields are provided with fingers that short out the waveguide formed by the amplifier enclosure, thereby reducing the tendency toward instability due to propagation of energy down the length of the enclosure.

The B+ and heater by-pass capacitor assemblies mount under the tube shields with the tubular capacitors beneath the vacuum tubes (see Fig. 3). The inductor forms are secured to the chassis by phosphor-bronze stampings which are soldered to silvered areas on the edges of the inductor forms. The same screws that secure the tube shields to the chassis simultaneously secure the capacitor assemblies and inductor clamps.

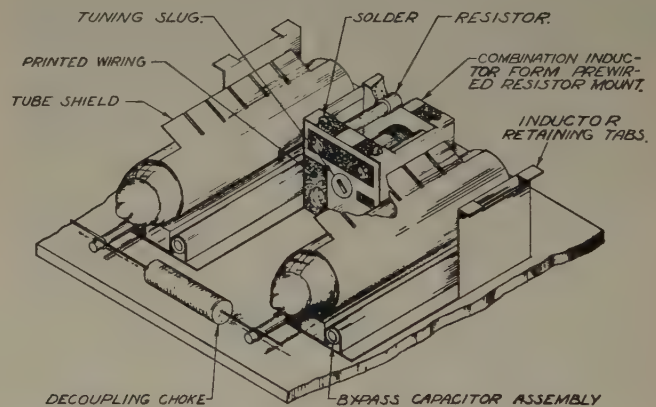


Fig. 3—Stage assembly of the Model II intermediate-frequency amplifier.

The inductors were wound with teflon-ceramic insulated wire and impregnated with silicone resin. Because this amplifier was sealed in a dry nitrogen atmosphere, no breakdown due to electrolytic action between the primaries and secondaries of these inductors is anticipated. The powdered-iron cores used to tune these inductors were adjusted from the inductors' low-potential ends. The physical properties, losses, and permeability of these cores change slightly after continuous operation at high temperatures, but the  $Q$  of the loaded network is sufficiently low so that variations in the loss and permeability of the slugs do not materially affect the bandwidth or tuning.

The use of ceramic for inductor forms made it possible to metallize a wiring pattern on their ends. Holes were provided in the ends of the inductor forms so that resistors could be inserted into these holes and soldered to the metallized wiring pattern (see Fig. 3).

The radio-frequency chokes were made from  $\frac{1}{8}$ -inch-diameter steatite tubing cut into 7/16-inch lengths. The ends were nicked and silvered. After firing, axial

leads were soldered into the ends and ceramic-insulated wire wound on each form.

The interstage components between the video amplifier and the cathode follower were mounted in a ceramic block. The ends of the ceramic block were silvered so that the components had only to be inserted into their respective holes and soldered into place to complete the circuit connections. The majority of the resistors in this amplifier were not required to dissipate any appreciable power. Several resistors that must dissipate considerable power were housed in the ceramic block to increase their power-dissipating capabilities. The resistors that have the most critical stability requirements are those used to load the tuned circuits. That these resistors are not called upon to dissipate appreciable power is a favorable factor.

Pretinned metal inserts were sweated into the center of each capacitor tube, making a very rigid structure. These metal inserts function as insulated tie-points for the heater and B+ chain decoupling chokes. The cathode and automatic-gain-control by-pass capacitors have tapped metal inserts so that they may be screwed vertically to the chassis with the inner metallized surface functioning as the grounded terminal. The heater and B+ by-pass capacitors have their outer metallized surface grounded. Since there is always the possibility that extreme stresses might cause the silver to separate from the ceramic base material, these capacitors were mechanically held in place by light-gauge metal stampings into which they were snapped before soldering. The stampings also function as grounding lugs.

All of the ceramic parts have been designed so that they may be extruded. Where formed and stamped metal pieces were required, these were inexpensively made of 0.005-inch sheet with the exception of the chassis plate. The chassis is die-stamped or perforated in a drill jig.

The interstage inductors are wound and the resistors inserted into the inductor forms in a preassembly operation. All hardware, preassembled interstage networks, by-pass capacitors, and tubes are mounted on the chassis, which is then set on edge in a cradle with the tube leads projecting upwards. It then progresses down a wiring line with the heater and B+ decoupling chokes mounted as a final wiring operation.

The basic layout of Model II permits many circuit modifications to be made without requiring radical modifications. There is sufficient room for larger inductors (for lower-frequency operation) to be incorporated without increasing the size of the amplifier. Because the by-pass capacitors have more capacity than is required for 60 Mc, they too will function quite satisfactorily at lower frequencies.

#### PRINTED CIRCUIT INTERMEDIATE-FREQUENCY AMPLIFIER

The printed-circuit-type intermediate-frequency amplifier is closely related in circuit and performance to

the model just described. The printed assemblies are contained in a hermetically sealed, rectangular metal case approximately  $6\frac{1}{2}$  inches long,  $2\frac{1}{4}$  inches wide, and  $\frac{5}{8}$  inch thick.

The design is based on a unit assembly of the principal parts associated with each electron tube. Each standard unit (see Fig. 4) is a tubular structure 2 inches long and  $\frac{1}{2}$  inch in diameter, enclosing a subminiature



Fig. 4—Model PC-IV, printed intermediate-frequency amplifier design based on unit construction of a single-stage unit. Unit printed assembly contains subminiature electron tube and principal circuit components associated with each stage, including inductor assembly. Three basic parts form the stage unit: (a) high-K ceramic tube, (b) steatite inductor form, and (c) high-K ceramic tube.

tube and an inductor assembly. As shown in Fig. 5, eleven of these cylindrical units are assembled to per-



Fig. 5—Front view of the complete assembly of the printed, miniature intermediate-frequency amplifier, Model PC-IV.

form the functions required in the intermediate-frequency amplifier. Eight of the units perform as tuned amplifiers and are stagger-tuned in pairs to give a band-

width of 10 Mc centered on 60 Mc. Supplementing these are a detector stage, a video amplifier, and a cathode-follower output stage.

All stages are similar in appearance and general construction with a common design evolved from a basic assembly of three ceramic parts and a subminiature vacuum tube. This assembly may be supplemented by the addition of decoupling chokes when necessary, as in the case of intermediate-frequency stages. Changes in the primary function of the assembly are accomplished by changing the pattern of the printed circuit and by using electron tubes of desired characteristics. Standardization of the construction of the assemblies was adopted. Such items as uniformity of tube connections and power-supply lead location facilitate construction, repair, and diversity of application of this type of assembly. In the amplifier as constructed, each stage unit is slipped into a silver-plated spring clip, which in turn is soldered to a single metal support. Individual units may be removed from or inserted into such an assembly by disconnecting the five interstage connections. It is believed that this feature, repair by stage replacement, may eliminate some previous objections to printed circuit devices where long service was mandatory.

The construction of the printed intermediate-frequency amplifier employs basic metallizing techniques for conductors and makes extensive use of ceramic materials, as is characteristic of the commercial printed-circuit activities of this country. Conductive patterns are made by imprinting the ceramic materials with a special silver-pigmented paint, using the screen-stencil processes. This simple mechanism of reproduction is readily adapted to the manufacture of printed circuit assemblies. The conductive paint is fired at temperatures of the order of 700°C to give it the required properties of good electrical conductivity, good adherence, and ability to accept solder.

The basic stage assembly is described in detail as follows: The first element is a high-K titanate tube approximately  $1\frac{1}{2}$  inches in length and 0.4-inch inside diameter so as to slip over the T-3 size subminiature tube. Assembled to one end of this long ceramic tube and the base of the subminiature tube is a steatite inductor form with a bifilar wound inductor. Over this inductor form is slipped the second tubular ceramic element which may be used to provide inductor shielding, additional by-passing capacity, or a circuit area for printed wiring of low distributed capacity by suitable selection of ceramic material and printed conductive pattern. Fig. 6 is a cutaway view of a typical stage showing the general construction of the parts and the details of the assembly. This figure also illustrates the multiple-layer construction of the long titanate tube. Metallized on the inside and with multiple capacitor plates printed on the outside, such a unit becomes a capacitor assembly with the capacitors produced *in situ* and by a single printing operation. In order to shield this assembly, a coating of

vitreous enamel is applied to the outside area of the tube and an all-over metallizing coat is then fired on top of the vitreous enamel.

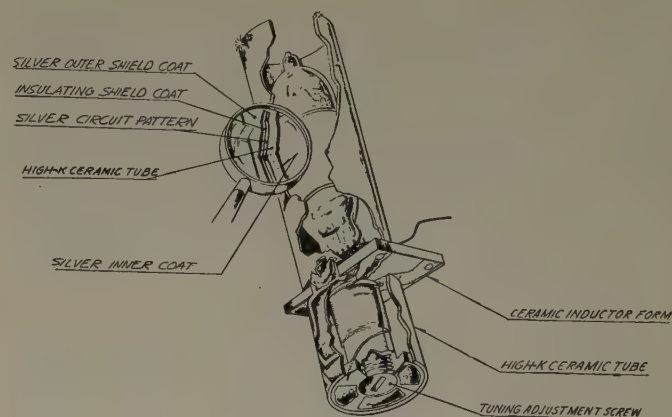


Fig. 6—Cutaway view of a typical stage of the Model PC-IV, printed-circuit amplifier.

An inductor assembly using wire-wound inductors was selected, for at the time of the conception of this design it was believed printed inductors could not compete with wire-wound inductors in extreme reduction of size. A special steatite spool was designed to function as an inductor form. The core is drilled and tapped with standard 8-32 screw threads. The inductor is thus made tunable by the use of either a powdered-iron screw or other standard metallic screw. Printed wiring on the rectangular flange on one end of this steatite form brings two internal connections outside of the stage assembly, making them easily accessible. In addition, the flange furnishes a convenient area for the location of a loading resistor with a minimum of distributed capacity. It also carries regularly spaced projections on its top surface, which serve to align the electron tube and its leads, and the long tubular capacitor assembly surrounding the electron tube. Another flange on the opposite end of the spool provides for termination of the "cold" end of the inductor windings and permits external connections to be readily made. This round flange is of a diameter which will permit the  $\frac{1}{2}$ -inch-long titanate tube to slip over it up to the rectangular flange and the point of juncture with the  $1\frac{1}{2}$ -inch-long tube. When printing the short tubes, it has been found expedient to reverse the order of ground connections used on the long tubes; the electrodes on the external surface, instead of the internal electrodes, are connected to ground. This arrangement makes it unnecessary to provide an additional shield coat, with its accompanying vitreous enamel layer, on the short tube.

The end of the  $1\frac{1}{2}$ -inch tube that joins the inductor assembly has eight small notches cut in its periphery, permitting the tube leads of the totally enclosed electron tube to be brought out of the assembly and appropriate connections made. The tube leads are preformed in a radial pattern, with the exception of the grid connection which joins directly to the inductor assembly as a very

short connection ( $\frac{1}{8}$  inch). The radial tube leads are aligned with the projections of the inductor form and protrude through the appropriate notches in the end of the  $1\frac{1}{2}$ -inch tube. The leads are then soldered directly to the proper capacitor or other connection. It is to be noted that there are only two internal connections in the stage unit and that these are made and verified before final assembly of the stage. All other connections are external, facilitating assembly, inspection, testing, disassembly, and repair. This system uses combination electrical and mechanical connections so that the operation of soldering joins the elements electrically and mechanically.

Orienting the circuit components around the circumference of the associated electron tube has been found to be very adaptable to variation in circuit requirements. For example, it almost always permits the optimum location, particularly with respect to lead length, of the major circuit element connected with a particular electron tube element. In the design of the intermediate-frequency amplifier, this adaptability permitted the standardization of the location of major circuit elements and connections.

In this basic design, leads as such are practically eliminated. As already pointed out, by-pass capacitors make up a large part of the total circuit elements. They are so located as to require short direct connections to the tube leads. Resistors usually do not require separate electrodes, but naturally fall so that they can use capacitor electrodes or existing "ground" patterns as their point of circuit contact. Heater and plate chokes do not require separate leads and are made to join directly to applicable circuit elements. As an example, the heater source is fed through a chain of chokes, each of which connects to the extreme ends of the printed heater by-pass capacitor electrodes which therefore also serve as conductors to supply the heaters with their normal voltage.

In addition to the elimination of many separate leads and connections, it can be seen that the design is also a step in the direction of the elimination of lumped constants. The design approaches a system of distributed constants contributing to operating efficiency and stability in some cases and, in particular, permitting the realization of economy in manufacture and fabrication as well as increasing the versatility of the basic concept in adapting it to many electronic circuits.

#### MODEL V

In the course of development of the Models II and PC-IV amplifiers, a request was made for a still smaller amplifier for a similar application. In order to obtain the desired shape, the configuration used in the printed-circuit model, with the inductors directly below the tubes, was selected. The mechanical design is one of several which were worked out to give the smallest possible size with available subminiature tube types. The particular advantage of this design is that it is built

up of stages mounted on individual chassis plates so that it lends itself to application in amplifiers having diverse electrical characteristics. Fig. 7 shows the stage detail

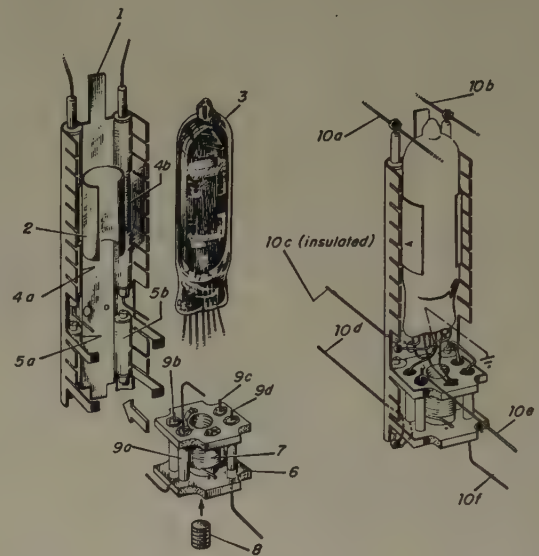


Fig. 7—Detail of stage construction used in the Model V amplifier. (1) Chassis plate. (2) Tube clip. (3) Tube. (4) Decoupling filter. (5) By-pass capacitors. (6) Inductor form. (7) Bifilar inductor. (8) Tuning core. (9) Resistors. (10) Interstage connections.

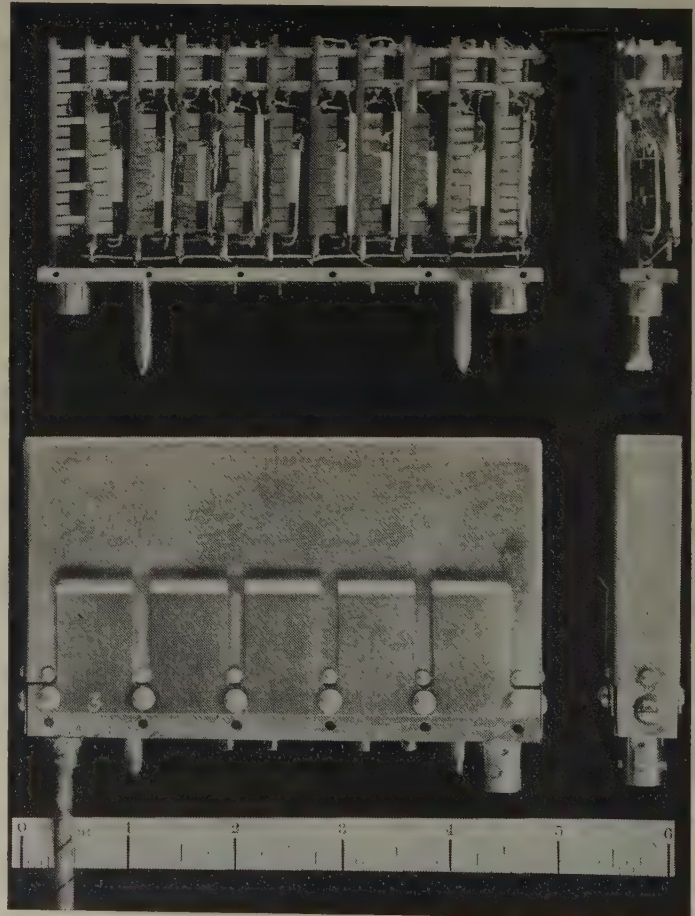


Fig. 8—Model V miniature intermediate-frequency amplifier. (Top) Side and end views of final assembly. (Bottom) Side and end views of encased amplifier in socket.

construction used in the Model V amplifier. While this stage is a grounded-grid amplifier used in the low-noise input circuit, its construction is typical of stages with other circuitry. It is shown to emphasize the fact that various types of circuitry may be used with this construction, although great attention to detail is required to utilize this advantage.

The assembly method consists of soldering the tubular capacitors, decoupling filters, and tube clip onto the chassis plate, then mounting the inductor form (already wound and with resistors in place) and the tube on the chassis plate. Connections between tube and inductor assembly are then made, after which the stages are assembled and interconnected. The various components shown in Fig. 7 are essentially the same as those used in the Model II, with modifications as required by the special design and circuitry.

A complete design was evolved to embody the identical circuitry of Fig. 1, as well as one for the special application. Fig. 8 shows the latter amplifier (ten tubes on separate chassis plates, plus one chassis plate con-

taining the input coupling unit) and the plug-in-type case with the socket assembly used for this amplifier. This assembly provides coaxial connections between amplifier and socket for input and output, complete shielding of all connections, and by-passing (1,000  $\mu\text{f}$  in socket) for power and gain-control leads.

A small production of the Model II has been completed and others are under way for the PC-IV and projected for the Model V. Some of the production models are being subjected to life tests at this time while others are intended for various other tests or installation application.

#### ACKNOWLEDGMENT

The authors wish to extend their appreciation to Benjamin L. Davis and P. V. Horton for their printed-resistor developments, and to Welfred M. Redler, Melvin E. Lutton, and Clinton O. Lindseth all of the Engineering Electronics Section, National Bureau of Standards, for their support in the development work on the various models.

## Interference Caused by More Than One Signal\*

RAYMOND M. WILMOTTE†, FELLOW, IRE

**Summary**—This paper covers a theoretical survey of the factors involved in establishing the effect on a desired signal of several interfering signals, including noise. It makes some suggestions toward the solution of the problems involved, and discusses the related problem of defining service area for broadcasting in the very-high and ultra-high frequencies.

### I. FACTORS AFFECTING THE MEASUREMENT OF INTERFERENCE

WHEN THERE IS more than one signal causing interference, the interfering signals will add to cause greater interference than is caused by any single one. The method by which they can be added depends on the following factors:

- A. The estimate of the median intensity of each signal;
- B. The time distribution of each signal;
- C. The correlation (if any) between the distributions of the several interfering signals;
- D. The probability distribution of the departure from the estimate of the median intensity;
- E. The type of interference (such as cochannel, adjacent channel, carrier beats, cross modulation, noise, and the like);

- F. The minimum acceptable ratio of desired to undesired signal for each type of interference;
- G. The threshold ratio of desired to undesired signal at which each type of interference becomes noticeable;
- H. The rate of change of objectionableness with signal intensity;
- I. The noise level and its distribution.

The factors will generally vary with the location so that in addition to time considerations a statistical analysis by location should be included.

A discussion on the estimate and effect of these factors follows.

#### *A. and B. The Estimate of Median Intensity and Time Distribution*

The true time distribution of each signal should be obtained by measurements over a long period; short time distributions are not satisfactory because the intensity variations are often caused by the combination of signals arriving over different paths. During a short interval of time, the variation in intensity is often largely controlled by the variation in the phase of two or more interfering signals. It is usually only over long periods that correct information can be obtained regarding the variations in the intensity. Such long periods are needed to take care of diurnal and seasonal variations, and, in some cases, variations from year to year.

\* Decimal classification: R171X430.11. Original manuscript received by the Institute, November 28, 1949; revised manuscript received, June 29, 1950.

† Raymond M. Wilmotte, Inc., Washington, D. C.

A factor which is very difficult to translate into quantitative terms is the nature of the time distribution of the interference. An interference that occurs 10 per cent of the time only may generally be considered acceptable. On the other hand, if it occurs one second out of every ten, it will not be acceptable, but if it occurs for one month out of every ten it will be. There is need for a factor representing the period between objectionable interferences. This might possibly be achieved by counting as five minutes of continuous interference any five-minute period in which objectionable interference occurs more than a specified number of times.

### C. Correlation Between Signal Distributions

Signals from two separate sources may be combined if their time distributions are known. Such estimates should strictly be divided into two parts, considering separately:

1. The parts of the time distributions which have no correlation with each other;
2. The parts of the time distributions which have some common relationship.

An example of what is being considered in this separation can be given from measurements of tropospheric waves. If measurements are made from two stations, there generally does not appear to be any correlation between the two, yet when each is analyzed separately there are found definite common diurnal and seasonal trends. These two trends should be removed from the time distribution, leaving only the uncorrelated time distribution. These can then be combined and the correlated trends reinstated to the combination to give the complete resultant time distribution. It is only when the correlation is considerable that it will become apparent in the final result.

### D. The Distribution of the Departure from the Median Values.

Over a period of time signals may vary from their median values so that when the median values of two signals are added together the procedure for obtaining the median value of the resultant is not the same as if the two signals were constant in value. The median value of the resultant depends on the distribution of the value of each of the signals and the law for calculating the resultant of several signals of fixed value. If this law is a simple addition (as in the case when the peak value of the resultant is the controlling function of interference) then the median value of the resultant is equal to the sum of the median values of the individual signals. If, however, the law is an rss (root-sum-square) law (as in the case when the total power is the controlling function of interference), then the times when a signal is strong is more than proportionately influential in affecting the value of the resultant, so that the median

value of the resultant will be greater than the rss of medians of the component signals.<sup>1</sup>

### E. Type of Interference

When several interfering signals on the same frequency which are not fading are received simultaneously, it is clear that, if they all carry the same modulation, the interference will be a function of the resultant of the signals. If the component signals are of constant intensity the resultant will vary depending on the phases of the components. The median amount of desired interference will be equal to that produced by a signal having an intensity equal to the median value of the resultant of the signals. That is equal to the root-sum-square of the rms intensities of the component signals.

This may be written in the form of the equation

$$R_{\text{med}}^2 = \sum_{r=1}^{r=n} E_r^2$$

where  $R_{\text{med}}$  is the median value of the equivalent single signal causing the same interference as the sum of the signals of intensity  $E_1, E_2, \dots, E_n$ .

This solution will not be accurate if the component signals carry different programs or in any way cause different types of interference. In television, for instance, there are several easily recognizable types of interference. Two synchronous signals carrying the same program will add correctly according to this formula if they arrive at the receiver without any substantial time delay.<sup>2</sup> However, it is possible to receive besides the desired signal a number of interfering signals; one may produce "Venetian blinds," another a background of an undesired program, and another the interference due to noncoincidence of the synchronizing pulses. Two or more of these types of interference to-

<sup>1</sup> This statement can be given in mathematical terms as follows: When the rss rule is applicable, it can be applied to a distribution curve only at the point corresponding to the rms (root-mean-square) value. Then

$$E_{\text{rms}}^2 = \sum_{r=1}^{r=n} E_{\text{rms},r}^2.$$

For percentages of time  $T$  shorter than that corresponding to the rms value

$$E_T^2 < \sum_{r=1}^{r=n} E_{T,r}^2$$

and for percentages of time longer than that corresponding to the rms value

$$E_T^2 > \sum_{r=1}^{r=n} E_{T,r}^2.$$

In the case of log normal distribution, it has been shown in the F.C.C. TID Report 4.2.1, "The Log-Normal Distribution" by Harry Fine, that for two equal signals

$$E_{\text{rms}} = E_{\text{median}} + 0.00532F^2$$

where  $E$  is in db and  $F = E_1/E_{99\text{db}}$ ,  $E_1$  and  $E_{99}$  being the signals exceeded 1 per cent and 99 per cent of the time, respectively. Figs. 1 and 2 show the trend of the probability curves when such equal signals as 2, 4, 8, and the like combine.

<sup>2</sup> If there is an appreciable time delay between them, they may act as though they produced quite different types of interference, so that in this case the above formula may not hold.

gether are more objectionable than one alone, but how much more? There is no experimental evidence providing an answer to this question. The problem seems similar to that of how to add oranges to elephants. It has been assumed commonly that the equivalent single signal producing the same degree of interference as several signals producing different types of interferences was equal to the root-sum-square of the product of the component signal and the minimum ratio of desired to undesired signal producing objectionable interference. This assumption is no more than a guess. It is unsupported by any experimental evidence, and, in fact, is unlikely to be correct, though in some cases it may be an adequate approximation.

In order to make correct estimates of interference, there is need of a unit for measuring objectionability. Such a unit, termed an "interference unit," is discussed below.

The lack of knowledge on the degree to which interference is objectionable, as far as allocation problems are concerned, does not exclude the possibility of making an estimate of the minimum percentage of time that

the interference is objectionable. Two broad concepts are of assistance. First, when the interferences are very strong the dominant interference is probably the only one that matters to the observer. Whether that is so or not, however, is not of interest for allocation purposes because under such conditions the reception is definitely unacceptable. The case of importance is the one in which the interference is on the borderline of being objectionable. Second, when interference from each component signal is weak, each being objectionable for only a small percentage of time ( $T_1, T_2 \dots$ ) two or more of them will seldom be objectionable at the same time unless there is a high degree of correlation between them. If there is no correlation, the total percentage of time  $T_0$  that at least one interfering signal produces objectionable interference is given by

$$T_0 = 100 - 100 \left(1 - \frac{T_1}{100}\right) \left(1 - \frac{T_2}{100}\right) \dots \left(1 - \frac{T_n}{100}\right) \\ \doteq \sum_{r=1}^{r=n} T_r.$$

The percentage of time  $T_0$  is lower than the total percentage of time that the interference is objectionable by an amount  $T_0'$  during which each component interfering signal is individually too low to be objectionable, yet when combined with the others becomes objectionable.

In our present state of knowledge, it is difficult to guess to what degree of accuracy  $T_0$  represents the percentage of time of objectionable interference, but in many practical cases it is likely to be a good approximation. The error  $T_0'$  is made up of periods when two or more signals are individually close to causing objectionable interference, for it is only then that combining several such interferences together may become objectionable.  $T_0'$  is therefore small when the percentage of time that a signal is close to being objectionable is small. That is likely to occur when the fading of the desired or undesired signals or both covers a wide range of intensity. It is less likely to occur, however, if a large change in signal intensity is required to produce a noticeable change in the objectionableness of the interference. While no quantitative values are available, it seems probable that some types of interference fall in this latter category.

When dealing with allocation, it is not so much the interference at one location that matters, but the interference at many locations, or more specifically the probability of interference with location. In this case, where probability with location is a consideration,  $T_0'$  will tend to be small if the signal intensity varies over a wide range from one location to another.

F. and G. The Minimum Acceptable and Threshold Ratio of Desired to Undesired Signals

Subjective tests, while they have not been carried out in detail seem to indicate that observers agree relatively closely on the value of the threshold point where

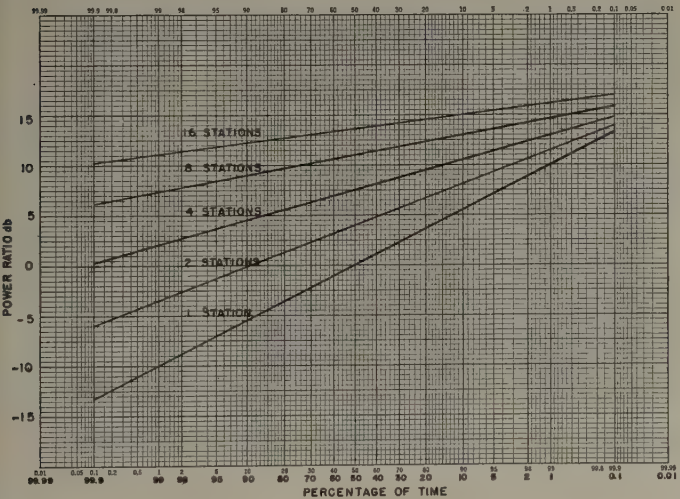


Fig. 1—Signal from several equal stations each having log normal distribution fading ratio 20 db.

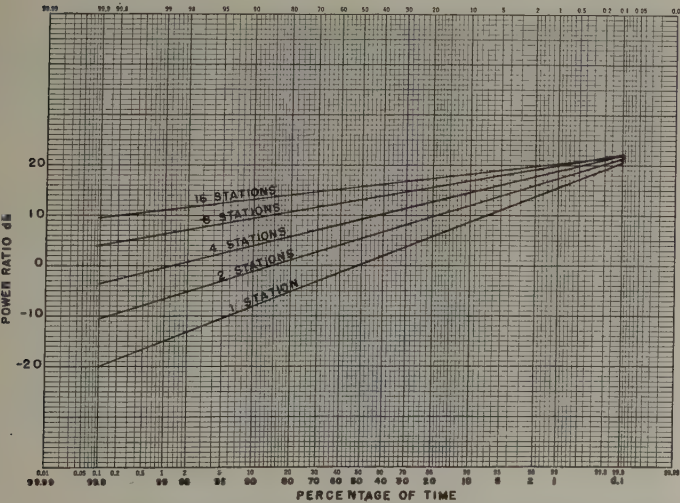


Fig. 2—Signal from several equal stations each having log normal distribution fading ratio 30 db.

interference just begins to be noticeable. The fact that they do indicates strongly that the threshold point is an important psychological quantity and justifies its use in formulas developed for evaluating total interference.

The value of the minimum acceptable ratio appears to vary over a wide range for different observers, and is therefore more difficult to estimate. It is, in part, subject to environment. A person who can only receive poor-quality signals will accept more interference than one who is accustomed to high-quality reception.

#### *H. The Rate of Change of Objectionableness with Signal Intensity*

The degree with which a type of interference varies with the desired- to undesired-signal ratio varies with the type of interference. Also, because of receiving-set characteristics as well as psychological effects, the signal intensity of the desired signal has some effect, but generally that is much less influential than the effect of the desired to undesired ratio.

#### *I. Noise Level*

Noise has usually been treated as a special kind of interference and in allocation problems it has usually been assumed that services may be rendered satisfactorily provided the signal is above a certain strength. This minimum signal strength has been estimated on the basis of the combination of the set noise, man-made noise, static noise, cosmic noise, and so forth. Actually, all these types of noise are subject to different treatment. For instance, set noise varies with each set and man-made noise varies with each location. The sum of various types of noises is therefore a statistical problem. There is no single figure which would be suitable for all locations, for all times, and for all sets. Noise should therefore be treated in the same way that other forms of interference are treated and the interference it causes added to other types of interference by the same kind of process.

It is incorrect to treat noise differently, as may be seen by considering the case of the interference caused by noise and an interfering signal from an undesired station. The combination of the two will produce more interference than each one separately. That result will be shown in the calculations if noise is treated as another interference, but it will not if it is treated, as has been customary, by merely requiring a certain minimum signal intensity to overcome it, and considering other interferences independently.

### II. THE INTERFERENCE UNIT

#### *A. Need for Unit*

In order to ascertain whether the degree to which an interference is objectionable varies over a wide range as the signal intensity varies, and how different types of interferences combine in terms of objectionableness, three things are necessary: first, to establish a unit, which might conveniently be called an "interference

unit"; second, to find how the interference varies with the ratio of desired to undesired signal and for different types of interference; third, how interference units for different types of interference may be added together.

A satisfactory definition of interference unit must be one which will provide a simple relationship between interference and signal intensity and which will also provide a simple formula for adding different types of interference. A suitable definition should make it possible to state that with a given ratio of desired to undesired signal of a certain type of interference the objectionableness is ( $x$ ) interference units. If another interfering signal of a different type exists which alone would produce ( $y$ ) interference units, then the total objectionableness would be a simple function of  $x$  and  $y$ , preferably  $(x+y)$  interference units. A prime requirement is that the same degree to which an interference is objectionable should be experienced with any type of interfering signal producing the same number of interference units.

The establishment of an interference unit would require subjective tests to establish the correlation between "objectionableness" and the ratio of desired to undesired signals of different types.

#### *B. Suggestions for Subjective Tests*

To carry out subjective tests in order to establish a formula for measuring the degree to which an interference is objectionable, it is desirable to establish a certain type of interference as a standard and compare other types with it. A convenient standard is an unmodulated carrier having a specified frequency difference with the carrier of the desired signal.

Having established a type standard, two identical receivers should be used. Both receivers carry the same program with one carrying a specified amount of the standard interference and the other carrying the interference under test. The program used should be of sufficient interest that the audience (which is now the detecting instrument) has a substantial degree of enjoyment in hearing or viewing it. Arrangements should be provided so that each member of the audience may select at will the reception for which the interference seems to him least objectionable. After a reasonable time each member of the audience will automatically select the program which he finds is least objectionable as far as the interference is concerned. The intensity of the interference is then changed in steps until fifty per cent of the audience finds one type of interference more satisfactory than the other. This test will provide information of the undesired to desired ratio of signal strength required to produce the same degree of interference as is produced by the standard interference. The interference under test can then be combined with the standard interference in various ratios, and the total disturbance compared with the standard. The result of this and similar tests will provide information on who interferences add together.

### C. Possible Definition of Interference Unit

Without a background of quantitative subjective tests it is probably futile to try to imagine what will be a suitable definition of an interference unit which meets the requirements of accuracy and simplicity. The suggestion made here for a definition is not to suggest the adoption of this unit, but to assist in the concept of it and thereby possibly help in the planning of subjective tests. The following train of reasoning leads to the definition given below.

It can readily be shown that the degree to which a signal is objectionable is not directly proportional to the ratio of undesired to desired signal. It can also be shown that for a certain ratio, a certain type of interference may be more objectionable than another, while for another ratio it may be less objectionable. The definition given below meets at least these qualitative requirements, but is given only as an example to explain the application of the use of an interference unit.

An interference which is below threshold is not noticeable. Hence, to add such an interference to one of another type already existing is not likely to increase the degree of interference. (Subjective tests may prove this to be incorrect when there are many different types of interference just below threshold.) This argument, if it is correct, leads to the suggestion that interference should use the threshold value as zero and be measured from that point up. Since most physiological reactions are logarithmic in character, it is therefore possible (again subject to tests so proving) that the degree to which an interference is objectionable may be measured as the quotient of two ratios; first, the ratio of desired to undesired signals, and second, that same ratio at the threshold point. This conclusion may be written mathematically as follows:

$$I = k(p - \sigma)$$

where  $I$  interference units are the "objectionableness" of the interference;  $p$  db is the ratio of the desired to undesired signal;  $\sigma$  db is the ratio of the desired to undesired signals at threshold and is dependent on the type of interference; and  $k$  is a constant dependent on the type of interference. If  $p_A$  is the minimum acceptable ratio of desired to undesired signal, then the minimum acceptable interference  $I_A$  is given by

$$I_A = k(p_A - \sigma).$$

For the standard type of interference,  $k=1$ . If  $\sigma_s$  is the ratio of desired to undesired signal at the threshold for the standard, and  $p_{As}$  the minimum acceptable ratio, then, if the above assumptions should prove to be sufficiently correct,

$$k = \frac{p_{As} - \sigma_s}{p_A - \sigma}.$$

Under those conditions the key values for any particular type of interference would therefore be  $p_A$  and  $\sigma$  to-

gether with the corresponding values for the standard type of interference.

### III. THE EFFECT OF ADDING INTERFERENCE ON THE AREA OR ON THE NUMBER OF PERSONS RECEIVING A SATISFACTORY SIGNAL

In many places in the United States, the service area of a station will be limited principally by the interference from other stations. The estimate of its service will depend greatly, therefore, on the estimate of the interference caused by several signals causing different types of interference. Their effect on the service area will also depend greatly on the definition of service area. In fact the definition of service area may have a great influence on the service rendered, particularly in areas saturated with stations.

The field-strength contour which is free from objectionable interference for a specified percentage of the time (90 per cent) has been taken in the Standards of Good Engineering Practice of the Federal Communications Commission as the limiting service contour of a station in the regular broadcast band. In the vhf and uhf bands there is no such contour, for on both sides of a contour there are persons receiving good signals for more than 90 per cent of the time. There are also on both sides persons receiving interference-free signals for less than 90 per cent of the time who find such signals acceptable. On both sides also there are persons receiving interference-free signals 80 per cent, possibly even 50 per cent, of the time and who find them acceptable.

It has been suggested that a contour be established as the limiting service contour where a specified percentage of the receiving locations (say 50 per cent) receive interference-free service not less than a specified percentage of the time (say 90 per cent). Any such definition of service tends to conceal some of the real facts. It tends to conceal the fact, for instance, explained in reference *H* to the Ad Hoc Committee Report, that the service deterioration is much more rapid when the service is limited by adjacent-channel interference than when it is limited by cochannel interference.<sup>3</sup> If a new station is installed causing interference, the number of persons affected is therefore not equal to the number of persons lying between lines representing the old and the new calculated service contours, for it will affect considerably many persons within the new service contour while some outside of it will not be affected as much. Actually, the number of persons affected will depend greatly on the type of interference.

It seems desirable, therefore, to be able to estimate the service area in terms of the number of persons or

<sup>3</sup> A graphical representation applicable to television is shown in Figs. 3 and 4 taken from reference *H* by R. M. Wilmotte and Harry Fine of the Ad Hoc Committee report. From the curves at any desired distance, the total percentage of locations receiving a specific grade of service or better can be read directly. Comparing Fig. 3, which refers to cochannel interference, and Fig. 4, which refers to adjacent-channel interference, shows clearly that in the case of adjacent-channel interference the service deteriorates with distance much more rapidly than in the case of cochannel interference.

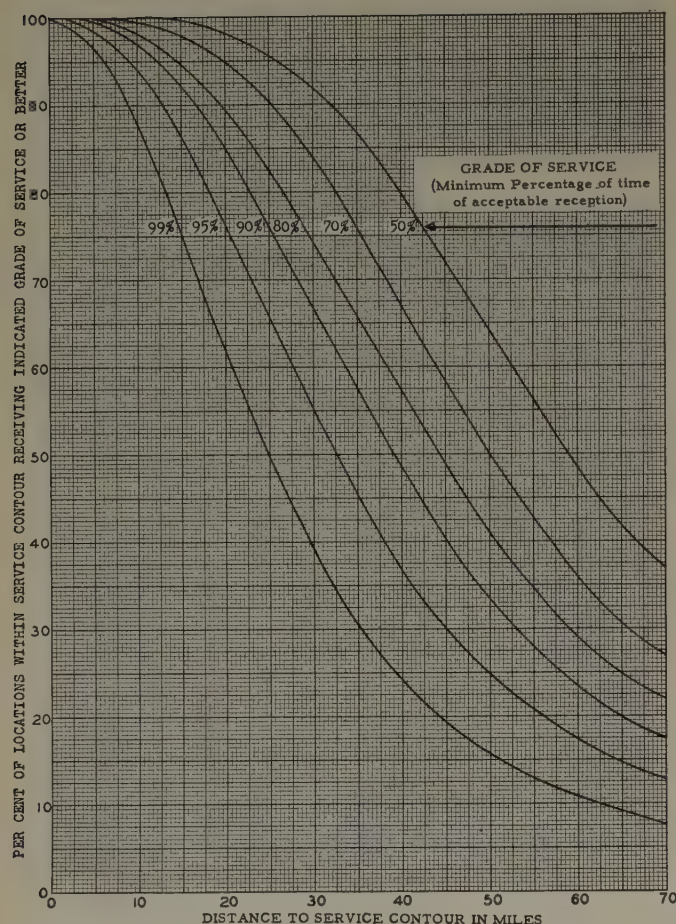


Fig. 3—Cochannel interference. Distance to service contour along radial between desired station and interfering station versus percentage of locations within service contour for various minimum grades of service. Frequency, 82 Mc; transmitting antenna heights, 500 feet; receiving antenna heights, 30 feet; equal power for both stations; station spacing, 150 miles; interference ratio, 40 db; type of area, rural.

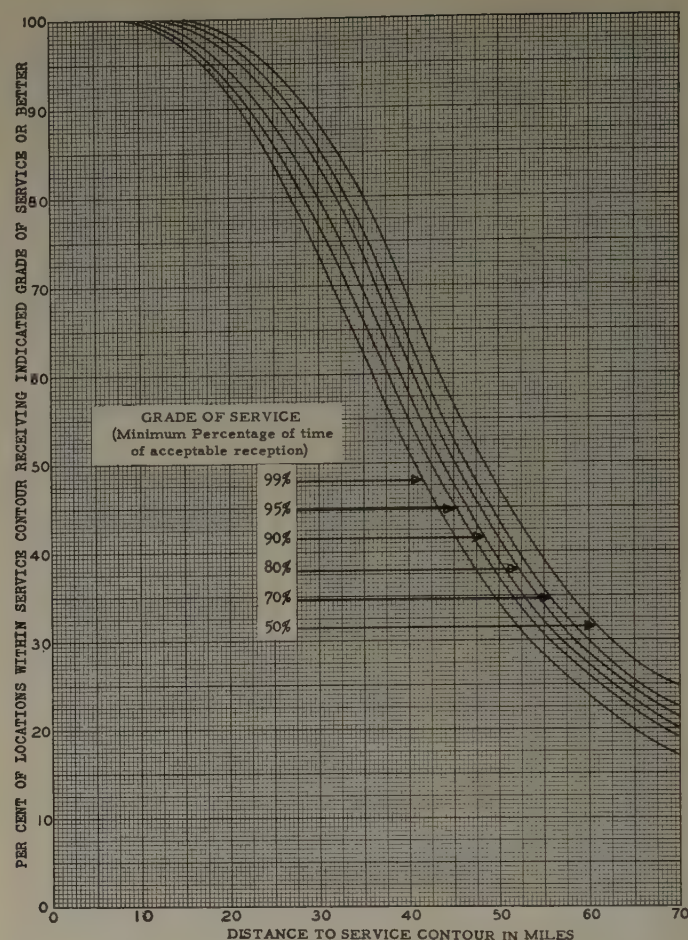


Fig. 4—Adjacent channel interference. Distance to service contour along radial between desired station and interfering station versus percentage of locations within service contour for various minimum grades of service. Frequency, 82 Mc; transmitting antenna heights, 500 feet; receiving antenna heights, 30 feet; equal power for both stations; station spacing, 75 miles; interference ratio, 6 db; type of area, rural.

receiving locations having a specified grade of service. The deterioration of the service due to a new source of interference could then be properly evaluated.

It would not be proper to find simply the total potential man hours of reception of interference-free signals, for clearly a person receiving a good signal only ten per cent of the time could hardly be considered as receiving a service at all. Ten-thousand such would not represent one-thousand man hours, they would represent more closely zero useful hours of reception.

It is suggested, therefore, that, in order to evaluate at least approximately the effect of an interfering signal, several grades of radio service be defined based only on the percentage of time of reception free from objectionable interference. (For instance, Grade A, 90 per cent or more of the time; Grade B, 70 per cent or more; and Grade C, 50 per cent or more. A minor modification of this may be considered as an alternative; namely, to

consider Grade A service as a service between 90 per cent and 100 per cent of the time; Grade B as service between 70 per cent and 90 per cent, and Grade C as service between 50 per cent and 70 per cent.) Inside a specified area served by vhf or uhf there will be persons receiving Grades A, B, or C service, or worse, depending on their location. The complete description of a service would then include the total area and the total number of persons expected to receive each grade of service.

#### ACKNOWLEDGMENTS

This paper in substantially this form was prepared for a subcommittee of the Ad Hoc Committee of the Federal Communications Commission. The author wishes to thank the members of the subcommittee, William Boese, Kenneth Norton, and Harry Fine, for helpful suggestions, particularly the latter who spent much time reading critically the original report.

# The $p$ -Germanium Transistor\*

W. G. PFANN† AND J. H. SCAFF‡

**Summary**—The transistor effect in  $p$ -type germanium is discussed and some properties are given for  $p$ -germanium transistors made in the laboratory. These exhibit higher cutoff frequency and somewhat lower current multiplication than their  $n$ -germanium counterparts. Under certain conditions a negative resistance "snap" effect is observed which is apparently peculiar to  $p$ -type germanium. Both types of transistor are governed by the same physical principles but they differ in the signs of the emitted carriers and of the bias voltages.

IN THE TRANSISTOR described by Bardeen and Brattain<sup>1,2</sup> two adjacent metallic points make rectifying contact to one surface of a small block of germanium. A third large-area contact comprises the base electrode. The point electrodes are designated emitter and collector and form the input and output terminals, respectively, of the transistor when connected as an amplifier. The basic material of the transistor described by Bardeen and Brattain is the electronic semiconductor,  $n$ -germanium.

The transistor effect has also been produced in the semiconductor  $p$ -germanium and a number of experimental  $p$ -germanium transistors have been made in the laboratory.<sup>3</sup> The preparation and properties of such  $p$ -germanium transistors are described herein, following a brief description of some characteristics of the semiconductors,  $n$ - and  $p$ -germanium.<sup>4</sup>

Germanium is an electronic semiconductor. Its electrical conductivity lies between those of metals and insulators. The conductivity can be increased by the addition of certain elements to the germanium. Such addition elements produce carriers of electrical charge which contribute directly to the conductivity. Addition elements fall into one of two classes, depending on the type of conductivity which they produce. Elements in Group V of the Periodic System donate conduction electrons to the germanium, thereby producing  $n$ -type ( $n$  for negative) conductivity. Elements of Group III accept an electron from the germanium lattice, leaving therein a net positive charge, or "hole," which can move freely through the lattice, producing  $p$ -type ( $p$  for positive) conductivity.<sup>5</sup> The existence of negative or positive carriers in semiconductors such as germanium can be demonstrated experimentally by measurement of the

Hall voltage, which is a transverse electromotive force produced by a magnetic field in a current-carrying conductor, the sign of the Hall voltage depending upon the sign of the moving carriers.

In actual practice, relatively pure  $n$ - or  $p$ -germanium, of resistivity on the order of 10 ohm-centimeters, has been used. Residual impurities may produce sufficient  $n$ -type conductivity for the requirements of transistors and high back-voltage rectifiers. Furthermore it has been shown by Scaff and Theuerer that such  $n$ -germanium can be thermally converted to  $p$ -germanium by heating at about 600°C or higher followed by quenching.<sup>6,7</sup> Such  $p$ -germanium, as well as  $p$ -germanium produced by additions of aluminum, have been used for the  $p$ -type transistors discussed herein.

While a good rectifying junction can be made between a metallic point and either  $n$ - or  $p$ -germanium, it should be noted that the polarity of the rectifying junction depends on the conductivity type of the semiconductor. Table I shows the polarity of the point with respect to the semiconductor in the two cases.

TABLE I

Conductivity Type of the Semiconductor	Polarity of Point	
	Forward Direction	Reverse Direction
$n$ type	+	—
$p$ type	—	+

In either type of transistor the collector is biased in the reverse direction and the emitter is usually biased in the forward direction. Operation of the  $n$ -germanium transistor may be explained on the basis that the current which passes through the emitter consists largely of holes, i.e., of carriers of opposite sign to those normally in excess in the body of the germanium. The holes are attracted to the negatively biased collector, so that a large part of the emitter current, introduced at low impedance, flows into the collector circuit and through a high impedance load, as may be seen in Fig. 1.

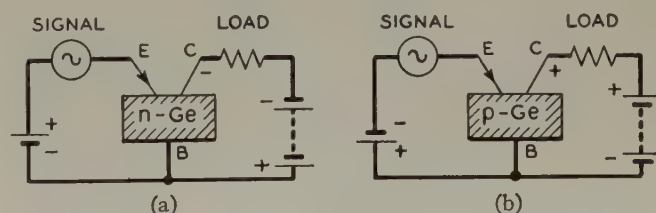


Fig. 1—Schematics showing operation of (a)  $n$ -germanium transistor, and (b)  $p$ -germanium transistor.

\* Decimal classification: R282.12. Original manuscript received by the Institute, May 25, 1950.

† Bell Telephone Laboratories, Inc., Murray Hill, N. J.

‡ J. Bardeen and W. H. Brattain, "The transistor, a semiconductor triode," *Phys. Rev.*, vol. 74, p. 230; July 15, 1948.

<sup>2</sup> J. Bardeen and W. H. Brattain, "Physical principles involved in transistor action," *Phys. Rev.*, vol. 75, pp. 1208-1225; April 15, 1949. And *Bell Sys. Tech. Jour.*, vol. 28, pp. 239-277; April, 1949.

<sup>3</sup> W. G. Pfann and J. H. Scaff, "The  $p$ -germanium transistor," *Phys. Rev.*, vol. 76, p. 459; August, 1949.

<sup>4</sup> For further information on electronic conduction in germanium, see footnote reference 2.

<sup>5</sup> J. H. Scaff, H. C. Theuerer, and E. E. Schumacher, " $P$ -type and  $N$ -type silicon and the formation of the photovoltaic barrier in silicon ingots," *Jour. Met.*, vol. 185, pp. 383-388; June, 1949.

<sup>6</sup> J. H. Scaff and H. C. Theuerer, "Preparation of High Back Voltage Germanium Rectifiers," National Defense Research Council, NDRC 14-555, October 24, 1945.

<sup>7</sup> H. C. Torrey and C. A. Whitmer, "Crystal Rectifiers," McGraw-Hill Book Co., New York, N. Y., chap. 12; 1948.

The result is a voltage gain and a power gain of an input signal. There may be an amplification of current as well.

In the *p*-germanium transistor a large part of the emitter current consists of electrons, which enter the germanium and are drawn to the positively biased collector. Voltage gain and current gain occur, just as in the *n*-type transistor.

The characteristics of the two types of transistor are such that in both, the output voltage is in phase with the input voltage for the grounded-base connection shown in Fig. 1. The input and output currents, however, are 180° out of phase in both types.

A transistor construction which has been found useful for laboratory studies is shown in Fig. 2. The germanium wafer assembly and the contact spring assembly are force-fitted into a metallic sleeve which serves also as the base electrode. The germanium surface is etched as has been described<sup>6,7</sup> for high back-voltage rectifiers, the etchant not being particularly critical. The con-

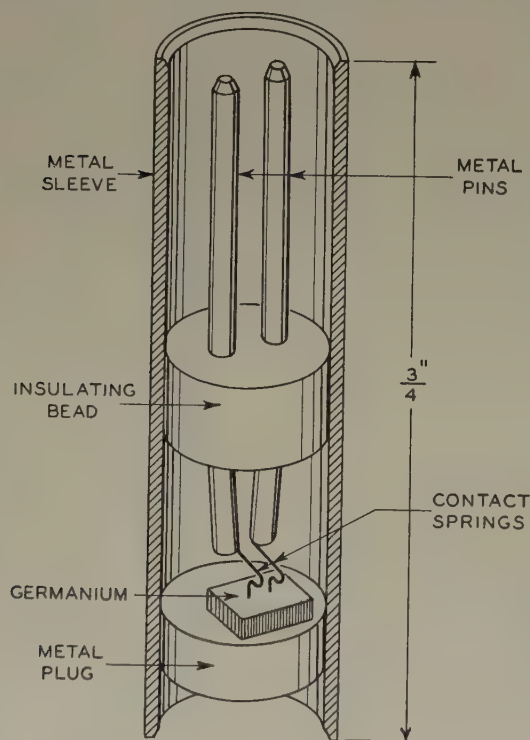


Fig. 2—Transistor construction used for laboratory studies.

tact springs are mounted on supporting pins which have previously been molded in a bakelite compound. The contact springs are phosphor bronze wires, 0.002 inch in diameter. Their tips are usually several mils apart; the wider spacing shown in Fig. 2 was used for clarity of illustration.

The ability of the emitter to inject electrons into *p*-germanium is greatly increased by an electrical forming treatment.<sup>8</sup> If desired, both point electrodes may

<sup>8</sup> J. Bardeen and W. G. Pfann, "Effects of electrical forming on the rectifying barriers of *n*- and *p*-germanium transistors," *Phys. Rev.*, vol. 77, pp. 401-402; December, 1950.

be pulsed, thereby enabling either point to be used as an emitter.<sup>9</sup>

The ac small signal performance of the *p*-germanium transistor at relatively low frequencies may be described with the help of the equivalent circuit of Fig. 3. The impedance associated with the emitter contact is  $r_e$ , and because the emitter junction is usually biased in the forward direction,  $r_e$  is small, of the order of hun-

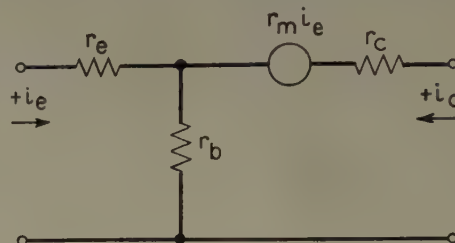


Fig. 3—Equivalent circuit representation of the transistor.

dreds of ohms. The impedance associated with the collector junction  $r_c$  is relatively high because the collector is biased in the reverse direction. The impedance in the semiconductor common to emitter and collector currents,  $r_b$ , represents a positive feedback. The transfer impedance  $r_m$  represents the active properties of the network. This network corresponds to the grounded-base connection.

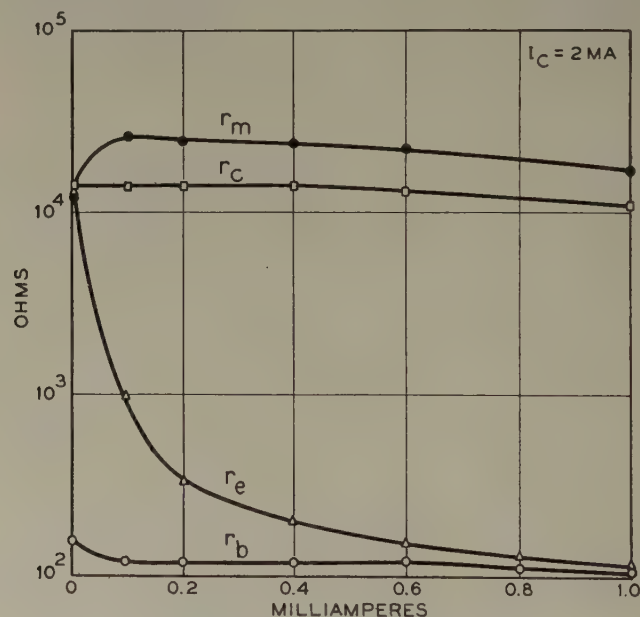


Fig. 4—Small-signal self- and transfer-impedances of a representative *p*-germanium transistor.

Some properties of a representative *p*-germanium transistor are shown in Fig. 4 in terms of these impedances. The  $r$ 's are plotted as functions of emitter current for a constant collector current of +2.0 milliamperes. At values of emitter current above a few tenths of a milliampere the impedances have the following approximate values:

$$\begin{array}{ll} r_e \sim 200 \text{ ohms} & r_c \sim 12,500 \text{ ohms} \\ r_b \sim 100 \text{ ohms} & r_m \sim 22,000 \text{ ohms} \end{array}$$

<sup>9</sup> W. G. Pfann, correspondence, "The transistor as a reversible amplifier," *Proc. I.R.E.*, this issue, p. 1222.

The ratio  $r_m/r_c$  is approximately equal to  $\alpha$ , the current multiplication factor and as may be seen,  $\alpha$  is greater than 1, usually being from about 1.5 to 2. The significance of  $\alpha > 1$  in terms of the physical operation of the transistor is that a given change in emitter current causes a larger change in the current through the collector.

As in the case of *n*-germanium, the gain of the *p*-germanium transistor falls off at frequencies in the megacycle range. Although we do not have an entirely complete comparison of the *n* and *p* devices, in which all quantities except frequency response are held constant, it does appear that the *p*-germanium transistor will operate at higher frequencies. Cutoff frequencies above 15 megacycles are common. (The cutoff frequency is that at which  $\alpha$  is down 6 db.) Measurements of cutoff frequency have been made by Rack and are plotted against distance between emitter and collector for both types of transistor in Fig. 5.

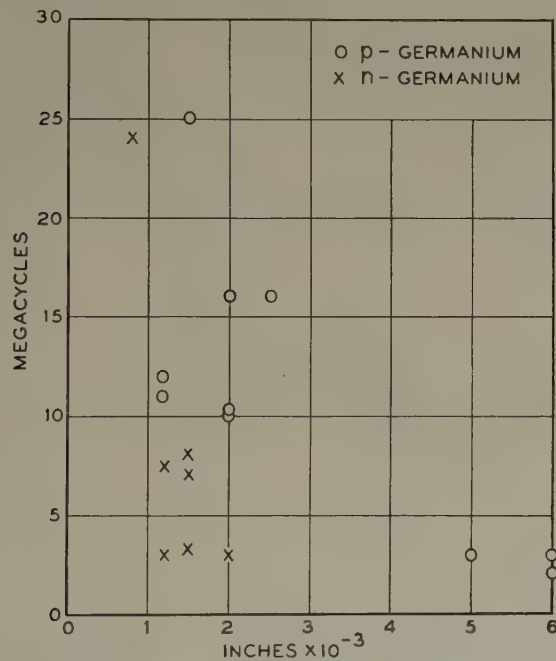


Fig. 5—Cutoff frequency of *n*-germanium and *p*-germanium transistors as a function of distance between emitter and collector.

An inherent feature of the *p*-germanium transistor may be at least partly responsible for its superiority in frequency response. It has been shown by recent measurements<sup>10</sup> that the mobility of electrons in germanium exceeds that of holes by a factor of about 1.5. Since the emitted carriers in the *p*-germanium transistor are electrons, and since it is dispersion in transit times of emitted carriers which limits the operating frequency, the observed superiority of the *p*-germanium transistor is at least partly to be expected.

The *p*-germanium transistor has a peculiarity which is found when the germanium is of particularly high resistivity. In such instances the forward current-volt-

age characteristic of the emitter has a negative resistance region of the voltage-maximum type, the peak voltage being on the order of several tenths of a volt. When the emitter bias exceeds this peak voltage, and if the series resistance in the emitter circuit is low, a sudden increase in  $I_e$  occurs which causes a corresponding change in the collector circuit. This action has been called the snap effect and may be seen in Fig. 6, which is a photograph of an oscilloscope trace in which the current-voltage characteristic of the collector junction is traced out at a frequency of 60 cycles per second, the voltage axis being horizontal, current axis vertical. The nearly horizontal portion of the trace represents the

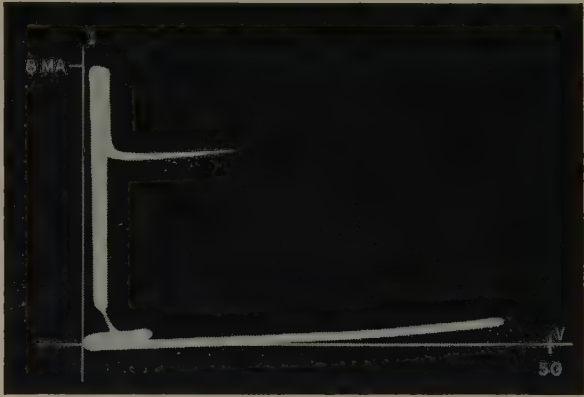


Fig. 6—Oscillograph pattern showing "snap" effect in *p*-germanium transistor.

reverse current through the collector for an emitter bias which is barely below the voltage maximum. At the extreme tip of the characteristic a small amount of positive feedback due to increase in collector current has caused the emitter bias to just exceed the voltage maximum. As a result the collector trace snaps suddenly to the high-current position as indicated by the vertical trace. The return trace remains in the low resistance position until the current falls to a small value, whereupon the characteristic again "snaps" back to the high-resistance position.

It is emphasized that the negative resistance in the forward current-voltage characteristic of the emitter is a diode effect, which can occur in a single point contact to *p*-germanium. It should not be confused with a similar effect which can be produced in an ordinary transistor by placing positive feedback resistance in the common base lead. However, the finite value of  $r_b$  does enter into the explanation of Fig. 6.

On the basis of a limited number of measurements there does not appear to be a significant difference in noise performance of *n*- and *p*-germanium transistors. Noise figures for *n*-germanium transistors are given in the literature.<sup>11</sup>

<sup>10</sup> W. Shockley, G. L. Pearson, and J. R. Haynes, "Hole injection in germanium," *Bell Sys. Tech. Jour.*, vol. 28, pp. 344-366; July, 1949.

<sup>11</sup> R. M. Ryder and R. J. Kircher, "Some circuit aspects of the transistor," *Bell Sys. Tech. Jour.*, vol. 28, pp. 367-400; July, 1949.

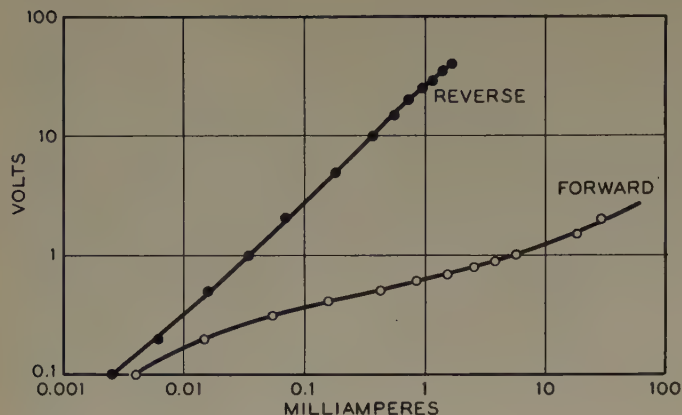


Fig. 7—Current-voltage characteristic of collector junction in a representative *p*-germanium transistor.

While it does not appear to have been recognized generally,<sup>12,13</sup> useful point contact rectifiers can be made with *p*-germanium. We have found that rectifier properties approaching those of high back-voltage *n*-germanium can be obtained by pulsing high-resistivity, thermally-converted *p*-germanium. The forward resistance before pulsing is high, on the order of thousands of ohms at one volt, but is greatly decreased by pulsing, to values on the order of 100 ohms or less. The reverse re-

<sup>12</sup> See p. 63 of footnote reference 7.

<sup>13</sup> W. C. Dunlap, Jr., and E. F. Hennelly, "Non-rectifying," *Phys. Rev.*, vol. 74, p. 976; October, 1948.

sistance is essentially unaffected by pulsing, and peak reverse voltages above 100 volts are common. A theoretical interpretation of electrical forming effects in *n*- and *p*-germanium has been made by Bardeen and Pfann.<sup>8</sup>

In Fig. 7 are shown the current-voltage characteristics of the collector junction in a representative *p*-germanium transistor at zero emitter current. Here, the reverse impedance is lower than the highest which can be obtained for it has been found, in common with the *n*-germanium transistor, that too high a reverse impedance is incompatible with best collector performance.

The attainment of transistor action in both *n*- and *p*-germanium provides further illustration of the dual nature of conduction processes in semiconductors such as germanium. In the *n*-type transistor, positive holes are injected into a semiconductor normally having an excess of electrons; in the *p*-type transistor, electrons are injected and the semiconductor normally has an excess of holes. In general, the properties of the *n*- and *p*-germanium transistors are similar, although in devices made up to the present time the *n*-germanium device appears to be superior in  $\alpha$  and inferior in frequency response to the *p*-germanium transistor. The present differences may be as much a result of the early status of transistor technology as of fundamental differences between the semiconductors themselves.

## Asymmetrically Driven Antennas and the Sleeve Dipole\*

RONOLD KING†, SENIOR MEMBER, IRE

**Summary**—The problem of determining the distribution of current in and the impedance of a cylindrical antenna asymmetrically driven by a discontinuity in scalar potential is formulated. An integral equation is derived and solved by the method of successive approximations to obtain general expressions for the current and the impedance. A simple approximate expression for the impedance of the asymmetrically driven antenna involving a series combination of the known impedances of symmetrically driven antennas is obtained. The impedance and the distribution of current for a cylindri-

cal antenna of length  $3\lambda_0/4$  driven  $\lambda_0/4$  from one end are evaluated. The broad-band properties are discussed.

Since the sleeve dipole with its image is equivalent to a superposition of two asymmetrically driven antennas, expressions for the impedance and distribution of current are obtained readily from the current distribution of the asymmetrically driven antenna. Both impedance and current distribution are determined for a sleeve dipole of over-all length  $3\lambda_0/4$  over a conducting plane driven  $\lambda_0/4$  from the plane. It is shown that this antenna has broad-band properties very superior to those of a conventional dipole.

### I. CURRENT AND IMPEDANCE FOR ASYMMETRICALLY DRIVEN CYLINDRICAL ANTENNAS

SINCE THE CENTER-DRIVEN cylindrical antenna is symmetrical with respect to its center, the currents in the halves are the same,  $I(-z) = I(z)$ ,

\* Decimal classification: R326.611. Original manuscript received by the Institute, February 21, 1950.

The research reported in this paper was made possible through support extended Cruft Laboratory, Harvard University, jointly by the Navy Department (Office of Naval Research), the Signal Corps of the U. S. Army, and the U. S. Air Force, under ONR Contract N5ori-76, T. O. 1.

† Cruft Laboratory, Harvard University, Cambridge, Mass.

and a single integral equation is involved. If the antenna is not center-driven, numerous complications arise. Most important is the fact that a transmission line feeding the antenna as in Fig. 1 is not in a neutral plane and is unbalanced by the different lengths of conductor attached to its ends. As a result, the transmission line is an important part of the radiating system since the co-directional components of current contribute significantly to the far-zone electromagnetic field. A radiating system which includes the transmission line as a radiating element is always undesirable. Obviously, the length and the location of the line should not play a part

in determining the field characteristics of an antenna system.

The only manner in which an asymmetrical antenna can be driven without introducing a radiating transmission line is by a generator completely within the

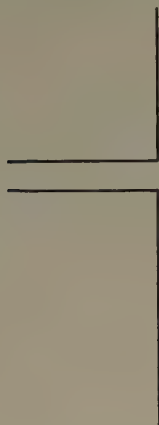


Fig. 1—Asymmetrical antenna driven by two-wire line.

metal surfaces of the antenna. Possible arrangements are those shown in Fig. 2. Except for projectiles, an-

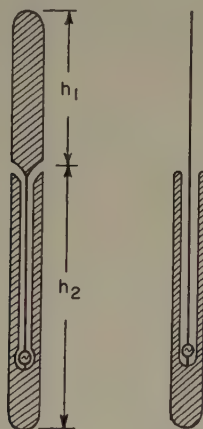


Fig. 2—Asymmetrical antenna internally driven.

tennas of this type are not practically important. However, a very useful antenna, the sleeve dipole, in which the radiating currents are equivalent to the superposition of the current in two asymmetrical driven antennas, is discussed later. Moreover, one step in the analysis of the collinear antenna<sup>1</sup> involves the determination of the current in an asymmetrical antenna center-driven by a discontinuity in scalar potential, i.e., by a slice generator.

The antenna to be analyzed is the thin cylindrical antenna drive off-center by a discontinuity in scalar potential at  $z=0$  as shown in Fig. 3. As in the analysis of the center-driven antenna, it is assumed that currents and charges are confined to the cylindrical envelope of radius  $a$ . Account may be taken of other surfaces at the

driving point if they exist just as for the center-driven cylinder. At the extremities, hemispherical ends approximate the idealized cylinder which has no chargeable surfaces other than on the cylinder itself. For simplicity, terms involving the surface impedance per unit length  $z^i$  are omitted so that the analysis applies strictly to a perfect conductor.

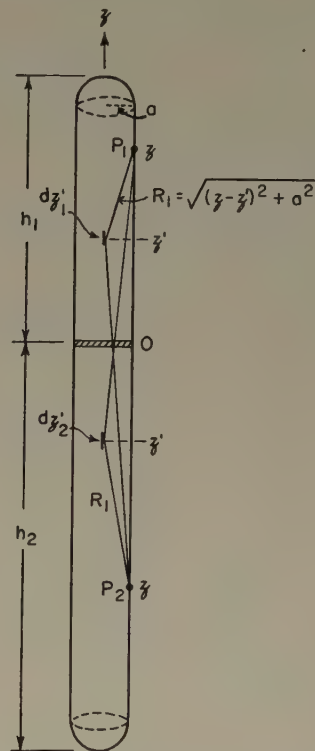


Fig. 3—Asymmetrically driven antenna to illustrate the notation.

Since the two parts of the antenna are not equal in length the distributions of current on them are necessarily different. This means that two currents,  $I_{1z}$  for  $(0 \leq z \leq h_1)$  and  $I_{2z}$  for  $(-h_2 \leq z \leq 0)$ , must be determined subject to the condition that they join smoothly at  $z=0$ ; that is

$$I_{10} = I_{20} = I_0. \quad (1)$$

The currents  $I_{1z}$  and  $I_{2z}$  are the unknowns in a pair of simultaneous integral equations. The equations are obtained, just as for the center-driven antenna,<sup>2</sup> by equating the Helmholtz integral for the vector potential to the solution of the differential equation for the vector potential. They are

$$\begin{aligned} 4\pi\nu_0 A_{1z} &= \int_0^{h_1} I_{1z'} K_1(z, z') dz' + \int_{-h_2}^0 I_{2z'} K_1(z, z') dz' \\ &= \frac{-j4\pi}{\zeta_0} [C_1 \cos \beta_0 z + C_2 \sin \beta_0 z]; \\ &\quad (0 \leq z \leq h_1) \end{aligned} \quad (2a)$$

<sup>1</sup> R. King, "Theory of Collinear Antennas," Cruft Laboratory Technical Report No. 91, October 3, 1949. To be published in *Jour. Appl. Phys.*

<sup>2</sup> R. King and D. Middleton, "The cylindrical antenna: current and impedance," *Quart. Appl. Math.*, vol. 3, p. 302; January, 1946.

$$4\pi\nu_o A_{2z} = \int_0^{h_1} I_{1z'} K_1(z, z') dz' + \int_{-h_2}^0 I_{2z'} K_1(z, z') dz' \\ = \frac{-j4\pi}{\zeta_o} [C_3 \cos \beta_o z + C_4 \sin \beta_o z]; \\ (-h \leq z \leq 0) \quad (2b)$$

where  $\beta_o = 2\pi/\lambda_o$ ,  $\zeta_o = 376.7$  ohms, and

$$K_1(z, z') \equiv \frac{e^{-j\beta_o R_1}}{R_1}; \quad R_1 \equiv \sqrt{(z' - z)^2 + a^2}. \quad (3)$$

These two equations must be solved for  $I_{1z}$  and  $I_{2z}$  subject to (1) and the conditions

$$I_{1z} = 0, \quad z = h_1; \quad I_{2z} = 0, \quad z = -h_2 \quad (4)$$

$$V_o = \lim_{\delta \rightarrow 0} [\phi_1(\delta) - \phi_2(-\delta)] = \phi_1(0) - \phi_2(0). \quad (5)$$

The scalar potential  $\phi$  is obtained as follows:

$$\phi_1(z) = \frac{j\omega}{\beta_o^2} \frac{\partial A_{1z}}{\partial z} = -C_1 \sin \beta_o z + C_2 \cos \beta_o z \quad (6)$$

$$\phi_2(z) = \frac{j\omega}{\beta_o^2} \frac{\partial A_{2z}}{\partial z} = -C_3 \sin \beta_o z + C_4 \cos \beta_o z. \quad (7)$$

Let

$$V_{10} \equiv \lim_{\delta \rightarrow 0} [\phi_1(\delta) - 0] = C_2; \\ V_{20} \equiv \lim_{\delta \rightarrow 0} [0 - \phi_2(-\delta)] = -C_4. \quad (8)$$

With (5)

$$V_o = \lim_{\delta \rightarrow 0} [\phi_1(\delta) - \phi_2(-\delta)] = V_{10} + V_{20} = C_2 - C_4. \quad (9)$$

The distribution functions for the currents may be introduced as follows:

$$f_1(z) \equiv I_{1z}/I_o; \quad f_2(z) \equiv I_{2z}/I_o \quad (10)$$

$$g_{11}(z, z') = f_1(z')/f_1(z); \quad g_{22}(z, z') = f_2(z')/f_2(z) \quad (11a)$$

$$g_{12}(z, z') = f_2(z')/f_1(z); \quad g_{21}(z, z') = f_1(z')/f_2(z). \quad (11b)$$

Let the following functions be defined:

$$\Psi_1(z) \equiv \int_0^{h_1} g_{11}(z, z') K_1(z, z') dz' \\ + \int_{-h_2}^0 g_{12}(z, z') K_1(z, z') dz'; \\ (0 \leq z \leq h_1) \quad (12a)$$

$$\Psi_2(z) \equiv \int_0^{h_1} g_{21}(z, z') K_1(z, z') dz' \\ + \int_{-h_2}^0 g_{22}(z, z') K_1(z, z') dz'; \\ (-h_2 \leq z \leq 0) \quad (12b)$$

$$\Pi_1(z) \equiv \int_0^{h_1} [I_{1z'} - I_{1z} g_{11}(z, z')] K_1(z, z') dz'$$

$$+ \int_{-h_2}^0 [I_{2z'} - I_{1z} g_{12}(z, z')] K_1(z, z') dz' \quad (13a)$$

$$\Pi_2(z) \equiv \int_0^{h_1} [I_{1z'} - I_{2z} g_{21}(z, z')] K_1(z, z') dz' \\ + \int_{-h_2}^0 [I_{2z'} - I_{2z} g_{22}(z, z')] K_1(z, z') dz'. \quad (13b)$$

The functions  $\Psi_1(z)$  and  $\Psi_2(z)$  are the ratios of vector potential to current at points  $z$  on the two parts of the antenna if  $f_1(z)$  and  $f_2(z)$  are the true distributions of current. These ratios are predominantly real and sensibly constant over each of the two parts of the antenna.<sup>2</sup> Therefore, let

$$\Psi_1(z) = \Psi_1 + \gamma_1(z); \quad \Psi_2(z) = \Psi_2 + \gamma_2(z) \quad (14)$$

where

$$\Psi_1 = |\Psi_1(z_r)|; \quad \Psi_2 = |\Psi_2(z_r)| \quad (15)$$

and  $z_r$  is a reference point on each part at which the magnitude of  $\Psi(z)$  approximates the mean constant value of  $|\Psi(z)|$ . The complex functions  $\gamma_1(z)$  and  $\gamma_2(z)$  are very small except near the respective ends at  $z = h_1$ ,  $z = -h_2$  where they become infinite, but  $I_z \gamma(z)$  vanishes.

With (10) to (15) and (9), (2) becomes

$$I_{1z} [\Psi_1 + \gamma_1(z)] + \Pi_1(z) \\ = \frac{-j4\pi}{\zeta_o} [C_1 \cos \beta_o z + V_{10} \sin \beta_o z]; \quad (0 \leq z \leq h_1) \quad (16a)$$

$$I_{2z} [\Psi_2 + \gamma_2(z)] + \Pi_2(z) \\ = \frac{-j4\pi}{\zeta_o} [C_3 \cos \beta_o z - V_{20} \sin \beta_o z]; \quad (-h_2 \leq z \leq 0). \quad (16b)$$

If the functions  $f_1(z)$  and  $f_2(z)$  were the true distributions of current, the difference integrals  $\Pi_1(z)$  and  $\Pi_2(z)$  necessarily would be zero. Since the true distributions are actually the unknowns, approximate distributions must be used, in which case  $\Pi_1(z)$  and  $\Pi_2(z)$  are the smaller the more closely the assumed distribution functions represent the true currents. It follows that the principal terms on the left are  $I_{1z}\Psi_1$  and  $I_{2z}\Psi_2$ . Hence

$$I_{1z} = \frac{-j4\pi}{\zeta_o \Psi_1} [C_1 \cos \beta_o z + V_{10} \sin \beta_o z] \\ - \frac{1}{\Psi_1} [\gamma_1(z) + \Pi_1(z)]; \quad (0 \leq z \leq h_1) \quad (17a)$$

$$I_{2z} = \frac{-j4\pi}{\zeta_o \Psi_2} [C_3 \cos \beta_o z - V_{20} \sin \beta_o z] \\ - \frac{1}{\Psi_2} [\gamma_2(z) + \Pi_2(z)]; \quad (-h_2 \leq z \leq 0). \quad (17b)$$

The simultaneous integral equations (17a) and (17b) involve no approximations not already contained in (2). They represent merely a rearrangement of the terms. However, the unknown currents  $I_{1z}$  and  $I_{2z}$  in (17a) and

(17b) occur in two types of terms, namely, the principal terms on the left and the difference integrals  $\Pi_1(z)$  and  $\Pi_2(z)$  on the right which are small if well-chosen distribution functions  $f_1(z)$  and  $f_2(z)$  are used. These equations may be solved by the same method of iteration used for the center-driven antenna.<sup>2</sup> The final formulas are as follows:

$$I_{1z} = \frac{j4\pi V_{10}}{\zeta_0 \Psi_1} \left\{ \frac{\sin \beta_0(h_1 - z) + M_1(h_1, z)/\Psi_1 + M_2(h_1, z)/\Psi_1^2 + \dots}{\cos \beta_0 h_1 + A_1(h_1)/\Psi_1 + A_2(h_1)/\Psi_1^2 + \dots} \right\} \quad (0 \leq z \leq h_1). \quad (18)$$

If the variable  $x$  is defined as follows:

$$x = -z, \quad (19)$$

the current  $I_{2z}$  in the lower part of the antenna has exactly the same form as (18) with appropriate changes in subscripts. Thus

$$I_{2z} = -I_{2x} = \frac{j4\pi V_{20}}{\zeta_0 \Psi_2} \left\{ \frac{\sin \beta_0(h_2 - x) + M_1(h_2, x)/\Psi_2 + M_2(h_2, x)/\Psi_2^2 + \dots}{\cos \beta_0 h_2 + A_1(h_2)/\Psi_2 + A_2(h_2)/\Psi_2^2 + \dots} \right\} \quad (0 \leq x \leq h_2). \quad (20)$$

$$Z_1 = \frac{V_1}{I_{10}} = \frac{V_{10}}{I_0} = \frac{-j\zeta_0 \Psi_1}{4\pi} \left\{ \frac{\cos \beta_0 h_1 + A_1(h_1)/\Psi_1 + A_2(h_1)/\Psi_1^2 + \dots}{\sin \beta_0 h_1 + B_1(h_1)/\Psi_1 + B_2(h_1)/\Psi_1^2 + \dots} \right\} \quad (25a)$$

$$Z_2 = \frac{V_2}{I_{20}} = \frac{V_{20}}{I_0} = \frac{-j\zeta_0 \Psi_2}{4\pi} \left\{ \frac{\cos \beta_0 h_2 + A_1(h_2)/\Psi_2 + A_2(h_2)/\Psi_2^2 + \dots}{\sin \beta_0 h_2 + B_1(h_2)/\Psi_2 + B_2(h_2)/\Psi_2^2 + \dots} \right\} \quad (25b)$$

Note that from (8)

$$V_{20} = \lim_{z \rightarrow 0} [0 - \phi_2(-z)] = \lim_{x \rightarrow 0} [0 - \phi_2(x)]. \quad (21)$$

Referred to  $x = -z$  this is the negative of  $V_1$  referred to  $z$ .

Since a solution of (18) in terms of  $z$  and  $h_1$  automatically gives a solution of (20) in terms  $x$  and  $h_2$ , it is necessary only to solve (18). In (18) the functions  $M_1$  and  $A_1$  are defined as follows:

$$M_1(h_1, z) = F_1(h_1, z) \sin \beta_0 h_1 - G_1(h_1, z) \cos \beta_0 h_1 + G_1(h_1, h_1) \cos \beta_0 z - F_1(h_1, h_1) \sin \beta_0 z \quad (22a)$$

$$A_1(h_1) = F_1(h_1, h_1). \quad (22b)$$

The functions  $F_1$  and  $G_1$  in (22a) and (22b) are defined as follows:

$$F_1(h_1, z) = F_{0z}(h_1) \Psi_1 - \int_0^{h_1} F_{0z'}(h_1) K_1(z, z') dz' - \int_{-h_2}^0 F_{0z'}(h_2) K_1(z, z') dz' \quad (23a)$$

$$G_1(h_1, z) = G_{0z}(h_1) \Psi_1 - \int_0^h G_{0z'}(h_1) K_1(z, z') dz' - \int_{-h}^0 G_{0z'}(h_2) K_1(z, z') dz'. \quad (23b)$$

The functions  $F_1(h_1, h_1)$  and  $G_1(h_1, h_1)$  are obtained from (23a) and (23b) with  $z = h_1$ . The zeroth-order function are

$$F_{0z}(h_1) = \cos \beta_0 z - \cos \beta_0 h_1; \quad G_{0z}(h_1) = \sin \beta_0 z - \sin \beta_0 h_1; \quad (0 \leq z \leq h_1) \quad (24a)$$

$$F_{0z}(h_2) = \cos \beta_0 z - \cos \beta_0 h_2; \quad G_{0z}(h_2) = -\sin \beta_0 z - \sin \beta_0 h_2; \quad (-h_2 \leq z \leq 0). \quad (24b)$$

The impedance of the asymmetrically driven antenna is obtained using (1) and (9), namely,  $I_{20} = I_{10} = I_0$ ;  $V_0 = V_1 + V_2$ , in conjunction with (18) with  $z = 0$  and (20) with  $x = 0$ . Thus, retaining terms through the second order, it follows from (18) and (20) that

where

$$B(h) \equiv M(h, 0). \quad (25c)$$

However, from (9) and (25a) and (25b)

$$V_0 = V_{10} + V_{20} = I_0(Z_1 + Z_2). \quad (26)$$

Since the impedance  $Z_0$  of the antenna is by definition  $Z_0 \equiv V_0/I_0$ , it follows, with (26), that

$$Z_0 \equiv V_0/I_0 = Z_1 + Z_2 \quad (27)$$

where  $Z_1$  and  $Z_2$  are given by (25a) and (25b). Therefore, the impedance of the asymmetrical antenna is a series combination of the impedances  $Z_1$  and  $Z_2$  characteristic of the individual parts when in each other's presence.

With  $Z_0$ ,  $Z_1$ , and  $Z_2$  defined in (25) through (27) it is possible to express the partial voltages  $V_{10}$  and  $V_{20}$  as follows:

$$V_{10} = V_0 \frac{Z_1}{Z_0}; \quad V_{20} = V_0 \frac{Z_2}{Z_0}. \quad (28)$$

Substitution of (28) in (18) and (19) gives the final expressions for current in terms of the driving discontinuity in scalar potential.

$$I_{1z} = \frac{j4\pi V_0}{\zeta_0 \Psi_1} \frac{Z_1}{Z_0} \left\{ \frac{\sin \beta_0(h_1 - z) + \sum_{i=1}^N M_i(h_1, z)/\Psi_1^i}{\cos \beta_0 h_1 + \sum_{i=1}^N A_i(h_1)/\Psi_1^i} \right\} \quad (0 \leq z \leq h_1) \quad (29a)$$

$$I_{2z} = \frac{j4\pi V_0}{\zeta_0 \Psi_2} \frac{Z_2}{Z_0} \left\{ \frac{\sin \beta_0(h_2 + z) + \sum_{i=1}^N M_i(h_2, -z)/\Psi_2^i}{\cos \beta_0 h_2 + \sum_{i=1}^N A_i(h_2)/\Psi_2^i} \right\} \quad (-h_2 \leq z \leq 0) \quad (29b)$$

$$Z_0 = Z_1 + Z_2 = \frac{-j\zeta_0}{4\pi} \left\{ \frac{\Psi_1 \left[ \cos \beta_0 h_1 + \sum_{i=1}^N A_i(h_1)/\Psi_1^i \right]}{\sin \beta_0 h_1 + \sum_{i=1}^N B_1(h_1)/\Psi_1^i} + \frac{\Psi_2 \left[ \cos \beta_0 h_2 + \sum_{i=1}^N A_i(h_2)/\Psi_2^i \right]}{\sin \beta_0 h_2 + \sum_{i=1}^N B_1(h_2)/\Psi_2^i} \right\} \quad (30)$$

This completes the general analysis of the asymmetrically driven antenna. It remains to express the solutions in terms of tabulated functions.

## II. FUNCTIONS AND PARAMETERS FOR ASYMMETRICALLY DRIVEN ANTENNAS

In order to evaluate the several integrals involved in the expressions (29) and (30) for the current in and impedance of an asymmetrically driven antenna, it is convenient to introduce the integrals

$$C_{11}(h, z) \equiv \int_0^h \cos \beta_0 z' K_1(z, z') dz' \quad (31a)$$

$$S_{11}(h, z) \equiv \int_0^h \sin \beta_0 z' K_1(z, z') dz' \quad (31b)$$

$$E_{11}(h, z) \equiv \int_0^h K_1(z, z') dz' \quad (31c)$$

where

$$K_1(z, z') \equiv \frac{e^{-j\beta_0 R_1}}{R_1}; \quad R_1 = \sqrt{(z - z')^2 + a^2}. \quad (32)$$

These integrals may be expressed in terms of the tabulated generalized sine and cosine integrals.<sup>3</sup>

Since

$$K_1(z, -z') = K_1(-z, z'), \quad (33)$$

it follows that

$$\int_{-h}^0 \cos \beta_0 z' K_1(z, z') dz'$$

<sup>3</sup> Staff of Computation Laboratory, "Tables of Generalized Sine and Cosine Integral Functions," vols. I and II, Harvard University Press, Cambridge, Mass.; 1949.

$$= \int_0^h \cos \beta_0 z' K_1(-z, z') dz' = C_{11}(h, -z) \quad (34a)$$

$$- \int_{-h}^0 \sin \beta_0 z' K_1(z, z') dz' = \int_0^h \sin \beta_0 z' K_1(-z, z') dz' = S_{11}(h, -z) \quad (34b)$$

$$\int_{-h}^0 K_1(z, z') dz' = \int_0^h K_1(-z, z') dz' = E_{11}(h, -z). \quad (34c)$$

Note that

$$C_{11}(h, z) + C_{11}(h, -z) = C_a(h, z) \quad (35a)$$

$$S_{11}(h, z) + S_{11}(h, -z) = S_a(h, z) \quad (35b)$$

where  $C_a(h, z)$  and  $S_a(h, z)$  are the functions occurring in the formulas for the center-driven antenna.<sup>2</sup>

With the notations (31) and (34), (23) becomes

$$F_1(h_1, z) = (\cos \beta_0 z - \cos \beta_0 h_1) \Psi_1 - C_{11}(h_1, z) + \cos \beta_0 h_1 E_{11}(h_1, z) - C_{11}(h_2, -z) + \cos \beta_0 h_2 E_{11}(h_2, -z); \quad (0 \leq z \leq h_1) \quad (36a)$$

$$F_1(h_1, h_1) = -C_{11}(h_1, h_1) + \cos \beta_0 h_1 E_{11}(h_1, h_1) - C_{11}(h_2, -h_1) + \cos \beta_0 h_2 E_{11}(h_2, -h_1) \quad (36b)$$

$$G_1(h_1, z) = (\sin \beta_0 z - \sin \beta_0 h_1) \Psi_1 - S_{11}(h_1, z) + \sin \beta_0 h_1 E_{11}(h_1, z) - S_{11}(h_2, -z) + \sin \beta_0 h_2 E_{11}(h_2, -z); \quad (0 \leq z \leq h_1) \quad (37a)$$

$$G_1(h_1, h_1) = -S_{11}(h_1, h_1) + \sin \beta_0 h_1 E_{11}(h_1, h_1) - S_{11}(h_2, -h_1) + \sin \beta_0 h_2 E_{11}(h_2, -h_1). \quad (37b)$$

In order to evaluate the expansion parameters  $\Psi_1$  and  $\Psi_2$  it is necessary to choose distribution functions for the currents in the two parts of the antenna. For this purpose the zeroth-order currents are good approximations. Therefore, let

$$f_1(z) \sim \sin \beta_0(h_1 - z); \quad f_2(z) \sim \sin \beta_0(h_2 + z) \quad (38a)$$

$$g_{11}(z) = \frac{\sin \beta_0(h_1 - z')}{\sin \beta_0(h_1 - z)}; \quad g_{22}(z) = \frac{\sin \beta_0(h_2 + z')}{\sin \beta_0(h_2 + z)} \quad (38b)$$

$$g_{12}(z) = \frac{\sin \beta_0(h_2 + z')}{\sin \beta_0(h_1 - z)}; \quad g_{21}(z) = \frac{\sin \beta_0(h_1 - z')}{\sin \beta_0(h_2 + z)}. \quad (38c)$$

With (38), (12a) becomes

$$\Psi_1(z) \sin \beta_0(h_1 - z) = \int_0^{h_1} \sin \beta_0(h_1 - z') K_1(z, z') dz' + \int_{-h_2}^0 \sin \beta_0(h_2 + z') K_1(z, z') dz' \quad (39a)$$

$$= \sin \beta_0 h_1 C_{11}(h_1, z) - \cos \beta_0 h_1 S_{11}(h, z) + \sin \beta_0 h_2 C_{11}(h_2, -z) - \cos \beta_0 h_2 S_{11}(h_2, -z); \quad (0 \leq z \leq h_1). \quad (39b)$$

The definition of  $\Psi_1 \equiv |\Psi_1(z_r)|$  involves the choice of a reference point  $z_r$  such that  $\Psi_1$  will be a good representation of  $|\Psi_1(z)|$  over its principal, essentially constant range. Since the contributions to  $\Psi_1(z)$  are predominantly from currents near the point  $z$ , it follows that currents in the lower part of the antenna, i.e., where  $-h_2 \leq z \leq 0$ , contribute negligibly to  $\Psi_1(z)$  defined at  $r=a$  on the upper part,  $0 \leq z \leq h_1$ , except in a small range near and at  $z=0$ . The continuity of current at  $z=0$ , namely,  $I_{20}=I_{10}=I_0$ , assures that significant contributions to  $\Psi_1(z)$  with  $z$  near zero by currents in the nearest section of the lower part of the antenna cannot differ appreciably from what they would be if  $h_2$  were equal to  $h_1$ . Therefore, the general distribution of  $\Psi_1(z)$  for the asymmetrically driven antenna must be essentially similar to  $\Psi_{K1}(z)$  for the center-driven antenna. Hence the same location of the reference point  $z_r$  must be appropriate. That is,

$$\Psi_1 = \begin{cases} |\Psi_1(0)|/\sin \beta_0 h & \text{for } \beta_0 h_1 \leq \pi/2 \\ |\Psi_1(h_1 - \lambda_0/4)| & \text{for } \beta_0 h_1 > \pi/2 \end{cases} \quad (40)$$

where  $\Psi_1(z)$  is given by (12a).

The parameter  $\Psi_2$  for the lower part of the antenna ( $-h_2 \leq z \leq 0$ ) is defined by (40) with (39) if subscripts 1 are replaced by 2, and 2 by 1 on  $h$  and  $\Psi$ , and  $-z$  is substituted for  $z$ .

With the functions  $F_1$  and  $G_1$  and the expansion parameters  $\Psi_1$  and  $\Psi_2$  reduced to forms involving only tabulated functions, a complete first-order solution for the current at all points in, and the driving-point impedance of, a cylindrical antenna of small cross section asymmetrically driven by a discontinuity in scalar potential is made available.

### III. APPROXIMATE IMPEDANCE OF ASYMMETRICALLY DRIVEN ANTENNA

The actual numerical evaluation of the general formulas for current and impedance given in Section II has not been carried out, since good approximations are readily obtained directly from the solution of the center-driven antenna for which numerical values are available.<sup>2</sup> The argument is straightforward.

The determination of the current in the upper part ( $0 \leq z \leq h_1$ ) of a cylindrical antenna driven at  $z=0$  involves the vector potential at all points on the surface of this part. Contributions to the vector potential arise from current in the entire antenna, but as has been pointed out before, these are significant only from currents very near the point  $z$  ( $0 \leq z \leq h$ ). Since  $z$  is confined to the upper part of the antenna in determining the current in this part, contributions by currents in the lower part are significant only in a small range when  $z$  is near zero. The actual magnitude of this range can be obtained directly from computations made by Winternitz<sup>4</sup>

to determine the significance of a gap of length  $2\delta$  at the center. In Winternitz's Figs. 9A to 9H, for example, is plotted  $|\Psi(z)| = |C^s(z)|$ , which is a magnitude proportional to the vector potential at  $r=a$  on the surface of the antenna and in the gap  $2\delta$ . Since the antenna is symmetrical, the value at  $z=0$ , the bottom of the dip, is due equally to currents in both halves of the antenna. Therefore the vector potential at  $z=0$  due to the lower half only is only one half the magnitude of this dip. Note that this is a value proportional to the vector potential due to the currents in the entire lower half at a distance  $\delta$  from the point of maximum current at  $z=-\delta$ . In Fig. 4 is a plot of the approximate variation of  $|\Psi(z)|$  (which is proportional to the vector potential) beyond the end of a section of conductor carrying a maximum current at that end due only to the current in that section. The insert shows  $|\Psi(z)|$  for half of an antenna and, framed on the right, the section beyond the end represented in the main diagram. The rate of decrease

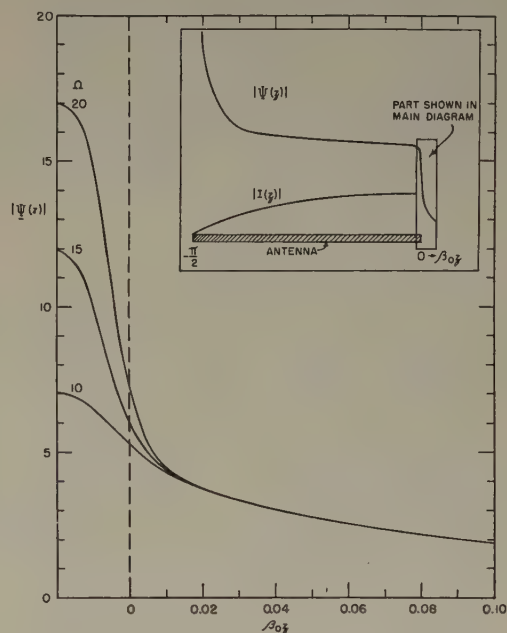


Fig. 4— $|\Psi(z)|$  as a function of  $\beta_0 z$  with  $\Omega$  as the parameter.

of  $|\Psi(z)|$  is such that at angular distances as small as 0.1 radians the magnitude is reduced to from 10 to 30 per cent, depending on the magnitude of  $\beta_0 a$  as specified by the value of  $\Omega$ . For small values of  $\beta_0 a$  the decrease is more rapid. Note that  $\beta_0 z = 0.1$  corresponding to  $z = 0.0159\lambda_0$ . Thus, the contribution to the vector potential even at  $z=0$  on an antenna by currents in the lower part ( $z < 0$ ) is confined to currents at distances not exceeding a tenth of a wavelength. Since rapid variations in the magnitude of the current in an antenna do not occur in such short distances, the requirement  $I_{10}=I_{20}=I_0$  is sufficient to assure that contributions to the vector potential at all points on the upper part of an

<sup>4</sup> T. W. Winternitz, "The Cylindrical Antenna Center-Driven by a Two-Wire Open Transmission Line," Cruft Laboratory Technical Report No. 56, August 15, 1948.

asymmetrically driven antenna, including  $z=0$ , by the actual currents in the lower part, do not differ effectively from what they would be if the distribution of current were the same in the lower part as in the upper, as when  $h_2=h_1$ . Evidently, currents in the lower part that are near enough to contribute to the vector potential in the upper part, i.e., near  $z=0$ , cannot differ appreciably from the currents in the adjacent upper part since  $I_{10}=I_{20}$ .

This means that in determining the distributions of current in the two parts, each may be analyzed separately as if each were a base-driven antenna over an infinite, perfectly conducting plane. That is,  $I_{1z}$  in (18) will be changed negligibly if  $h_2$  is set equal to  $h_1$  wherever it occurs in  $\Psi$  and in the  $F$  and  $G$  functions that contribute to the  $M$  and  $A$  functions. Similarly,  $I_{2z}$  may be determined from (20) by setting  $h_1$  equal to  $h_2$  in the corresponding integrals. Each of these problems is equivalent to the center-driven antenna.<sup>2</sup> The current  $I_{1z}$  in the upper part is the same as in the upper part of a center-driven antenna of half-length  $h_1$ ; the impedance  $Z_1 = V_{10}/I_{10}$  is one half that of the center-driven antenna. Similarly,  $I_{2z}$  in the lower part is the same as that in the lower part of a center-driven antenna of half-length  $h_2$ , and the impedance  $Z_2 = V_{20}/I_{20}$  is one half that of the center-driven antenna. The impedance  $Z_0$  of the asymmetrical array is the series combination of  $Z_1$  and  $Z_2$ .

$$Z_0 = Z_1 + Z_2. \quad (41)$$

The currents in the two isolated antennas of half-length  $h_1$  and  $h_2$ , respectively, are related to each other and to the asymmetrical antenna they are to represent by setting

$$V_{10} = V_0 Z_1 / Z_0; \quad V_{20} = V_0 Z_2 / Z_0. \quad (42)$$

Hence, with  $Y = 1/Z$ ,

$I_{1z}$ (upper part of asymmetrical antenna)

$$= \frac{Y_0}{Y_1} I_{1z}(\text{upper half of symmetrical antenna of half-length } h_1) \quad (43a)$$

$I_{2z}$ (lower part of asymmetrical antenna)

$$= \frac{Y_0}{Y_2} I_{2z}(\text{lower half of symmetrical antenna of half-length } h_2) \quad (43b)$$

where

$$Y_1 = 2(Y_0 \text{ of symmetrical antenna of half-length } h_1) \quad (44)$$

$$Y_2 = 2(Y_0 \text{ of symmetrical antenna of half-length } h_2) \quad (45)$$

$$Y_0 = \frac{Y_1 Y_2}{Y_1 + Y_2}. \quad (46)$$

As a numerical example consider the asymmetrical antenna defined as follows:

$$\beta_0 h_1 = \pi/2; \quad \beta_0 h_2 = \pi \quad (47a)$$

$$\Omega_1 = 2\ln(2h_1/a_1) = 10; \quad \Omega_2 = 2\ln(2h_2/a_2) = 10. \quad (47b)$$

Since  $h_2=2h_1$ ,  $a_2=2a_1$ . The second-order King-Middleton<sup>2</sup> admittances are

$$\begin{aligned} \beta_0 h_1 = \pi/2: \quad Y_1 &= 2(9.56 - j4.62) \times 10^{-3} \\ &= (19.12 - j9.24) \times 10^{-3} \\ &= 23 \times 10^{-3} / -23.2^\circ \text{ mhos} \end{aligned} \quad (48a)$$

$$\begin{aligned} \beta_0 h_2 = \pi: \quad Y_2 &= 2(1.0 + j1.8) \times 10^{-3} \\ &= (2 + j3.6) \times 10^{-3} \\ &= 4.12 \times 10^{-3} / 61^\circ \text{ mhos.} \end{aligned} \quad (48b)$$

The input admittance of the asymmetrical antenna is

$$Y_0 = (2.51 + j3.1) \times 10^{-3} = 3.98 \times 10^{-3} / 51^\circ \text{ mhos.} \quad (48c)$$

The input impedance is

$$Z_0 = 158 - j195 = 251 / -51^\circ \text{ ohms.} \quad (48d)$$

The input current of the asymmetrical antenna is

$$\begin{aligned} \frac{I_0}{V_0} &= 2.51 + j3.1 \\ &= 3.98 / 51^\circ \text{ milliamperes per volt.} \end{aligned} \quad (49)$$

The distribution of current along the individual parts is the same as if this part were half of a center-driven antenna, but multiplied for part 1 ( $\beta_0 h_1 = \pi/2$ ) by  $Y_0/Y_1$ , for part 2 ( $\beta_0 h_2 = \pi$ ) by  $Y_0/Y_2$ . Numerical values for the special case under consideration are

$$\frac{2Y_0}{Y_1} = 0.346 / 74.2^\circ; \quad \frac{2Y_0}{Y_2} = 1.936 / -10^\circ. \quad (50)$$

The factor 2 before  $Y_0$  is necessary for comparison with the center-driven antennas which have admittance  $Y_1/2$  and  $Y_2/2$ , respectively. Accordingly, the distributions of the magnitudes and phases of the currents per unit driving voltage are

$$\begin{aligned} \left| \frac{I_{1z}}{V_0} \right| &= 0.346 \left| \frac{I_z}{V_0} \right|_{\beta_0 h = \pi/2}; \\ \theta_{1z} &= \theta_{I, \beta_0 h = \pi/2} + 74.2^\circ \end{aligned} \quad (51a)$$

$$\begin{aligned} \left| \frac{I_{2z}}{V_0} \right| &= 1.936 \left| \frac{I_z}{V_0} \right|_{\beta_0 h = \pi}; \\ \theta_{2z} &= \theta_{I, \beta_0 h = \pi} - 10^\circ. \end{aligned} \quad (51b)$$

Curves showing the distributions of the magnitude and phase of the current referred to the driving voltage for the asymmetrical antenna with  $h_1 = \lambda_0/4$ ,  $h_2 = \lambda_0/2$  are in Fig. 5. These were obtained from approximate second-order curves of the King-Middleton distribution of current.<sup>5</sup>

<sup>5</sup> The second-order current curves will be included in a subsequent technical report. First-order curves are given in footnote reference 2.

Since the input impedance of the asymmetrical antenna is approximately equal to the sum of the impedances of two antennas of different and quite arbitrary lengths, it is interesting to consider the possibility of selecting these lengths so that the sum of the two im-

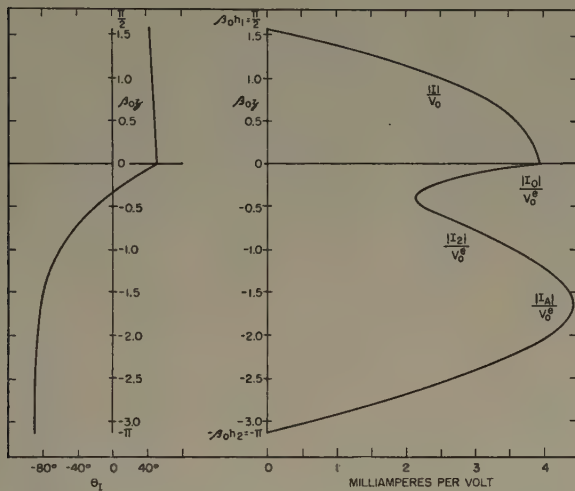


Fig. 5—Current distribution on asymmetrical antenna.

pedances is less sensitive to changes in  $\beta_0 h = \omega h/v_0$  than either one separately; that is, is it possible to select  $h_1$  and  $h_2$  so that the asymmetrically driven antenna has desirable broad-band characteristics? The answer is in the affirmative if it is possible to design the two parts 1 and 2 of the asymmetrical antenna so that for a given change in frequency from  $\omega_0 - \Delta\omega$  to  $\omega_0 + \Delta\omega$  the resulting change in  $R_1$  and  $X_1$  is compensated by a nearly equal and opposite change in  $R_2$  and  $X_2$ . For frequency range from  $\omega$  to  $\omega + \Delta\omega$ ,  $\beta_0 h_1 = \omega h_1/v_0$  increases to  $(\omega + \Delta\omega)h_1/v_0$ , and  $\beta_0 h_2 = \omega h_2/v_0$  increases to  $(\omega + \Delta\omega)h_2/v_0$ . Thus, the change in  $\beta_0 h$  is  $\Delta\omega h_1/v_0$ , and the change in  $\beta_0 h_2$  is  $\Delta\omega h_2/v_0$ . Evidently, the increase in  $\beta_0 h_2$  is greater than that in  $\beta_0 h_1$  if  $h_2$  is greater than  $h_1$ , so that the slope of the  $R_2$ ,  $X_2$  curves should be smaller than, and of sign opposite to, the slope of  $R_1$ ,  $X_1$  curves. Since for all moderately thin antennas the regions of negative slope for  $X$  are always much steeper than regions of positive slope, it is difficult to satisfy these conditions without using sections that are several half wavelengths long. If both  $h_1$  and  $h_2$  are to be under a wavelength and  $\Omega_1$  and  $\Omega_2$  are not to differ greatly, a reasonably broad-band antenna can be constructed by choosing  $h_1$  to maximize  $X_1$  and  $h_2$  to minimize  $X_2$  at  $\omega = \omega_0$ . For  $\Omega_1 = \Omega_2 = 10$ ,  $h_1 = 2.2v_0/\omega_0$  and  $h_2 = 2.9v_0/\omega_0$  where  $v_0 = 3 \times 10^8$  m per second. A range of  $\omega$  defined by

$$0.9\omega_0 \leq \omega \leq 1.1\omega_0 \quad (52a)$$

corresponds to

$$1.98 \leq \beta_0 h_1 \leq 2.42; \quad 2.61 \leq \beta_0 h_2 \leq 3.19. \quad (52b)$$

Curves showing the variation of  $R_1$  and  $X_1$ ,  $R_2$  and  $X_2$ ,  $R_0 = R_1 + R_2$ , and  $X_0 = X_1 + X_2$  are shown in Fig. 6 as a function of  $\omega$  in the range from  $0.9\omega_0$  to  $1.1\omega_0$ . It is seen that over this 20-per cent frequency change,  $R_0$  and  $X_0$  are reasonably constant. It is to be noted that with a corresponding choice of  $h_1$  and  $h_2$  for thicker antennas for which  $\Omega_1$  and  $\Omega_2$  are smaller than 10, a considerable extension of this approximately constant range is possible.

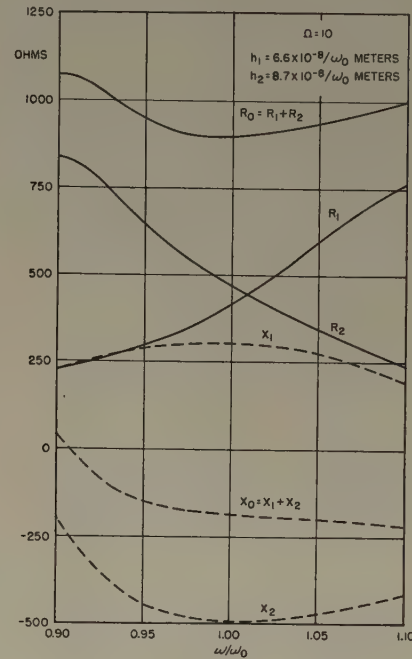


Fig. 6—Broad-band characteristics of asymmetrical antenna.

#### IV. THEORY OF THE SLEEVE DIPOLE

An antenna of considerable importance because of its broad-band properties is the *sleeve dipole* shown in Figs. 7(a) and 7(b). It consists essentially of the vertically extended inner conductor and outer conductor of a

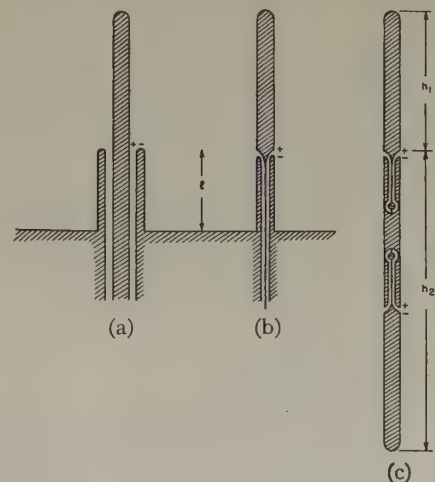


Fig. 7—Sleeve dipole driven by a coaxial line over an image plane and its equivalent symmetrical structure.

coaxial line over a horizontal conducting plane. It differs from the conventional base-driven half-dipole over a conducting plane in that the sheath of the coaxial line does not end at this plane but extends above it a distance  $l$ . This is equivalent to moving the gap across which a voltage is maintained by the feeding transmission line upward from  $z=0$  at the conducting plane to  $z=l$ . Application of the theorem of images to obtain an equivalent symmetrical structure yields the internally driven antenna of Fig. 7(c) in which two generators maintain equal voltage respectively across two gaps at  $z=\pm l$ .

Owing to the linearity of the Maxwell equations, the principle of superposition is applicable, so that the current at any point along the antenna in Fig. 7(c) is the algebraic sum of the currents maintained independently by the generators. This is illustrated schematically in Fig. 8.

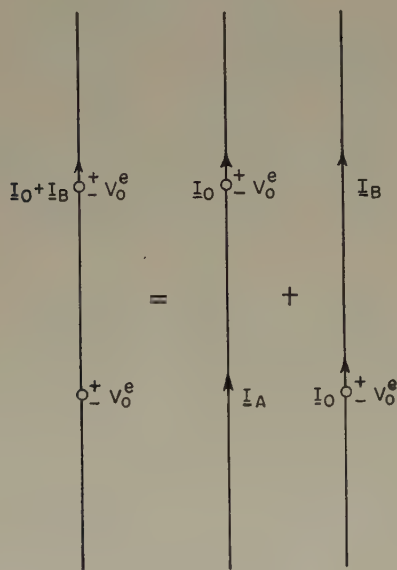


Fig. 8—Decomposition of the symmetrically driven antenna into two asymmetrically driven antennas.

The determination of the currents due to the upper generator in the symmetrical antenna on the left in Fig. 8 reduces to the analysis of the current in the middle antenna in Fig. 8. Since this is the asymmetrically driven antenna already analyzed, this part of the problem is solved. The current in the antenna on the right in Fig. 8 may be obtained directly from that in the middle antenna by interchanging ends. Finally, the resultant current in the antenna on the left is the algebraic sum of the currents in the other two.

The impedance at the terminals of the upper generator in the antenna on the left in Fig. 8 is given by the driving voltage  $V_o^e$  divided by the total current in the terminals. Thus,

$$Z_{in} = V_o^e / (I_o + I_B) = V_o^e / (I_o + I_A). \quad (53a)$$

The admittance is

$$Y_{in} = (I_o + I_A) / V_o^e. \quad (53b)$$

The numerical evaluation of the impedance and the distribution of current in a sleeve dipole is complicated by several factors. Consider first the simplest case of the symmetrical antenna in Fig. 9(a) where the driving voltages are maintained by discontinuities in scalar potential. To permit numerical evaluation, let  $\beta_o h_1 = \pi/2$ ,  $\beta_o h_2 = \pi$ , as shown. Also let  $h_2/a_2 = 75$  so that  $\Omega_2 = 2 \ln(2h_2/a_2) = 10$ . Since  $h_2 = 2h_1$ , it follows that with  $a_2 = a_1$ ,  $\Omega_1 = 2 \ln(2h_1/a_1) = 9.3$ . Since the impedance of an antenna of half length  $h_1 = \lambda_o/4$  varies only slowly with  $\Omega$ , an error of only a few per cent is made if it is assumed for simplicity that  $\Omega_1 = 10$ . This value corre-

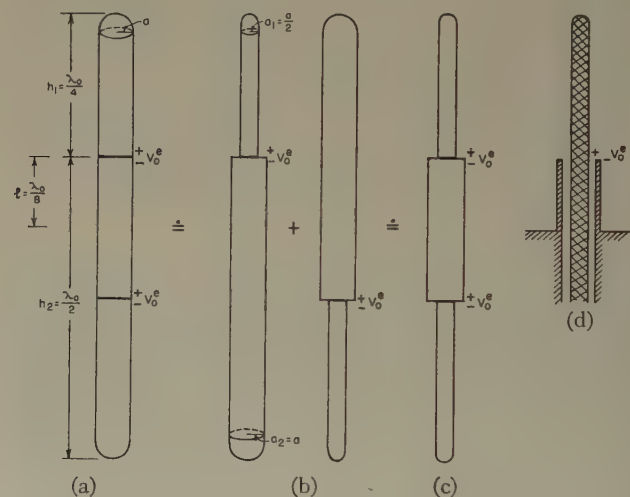


Fig. 9—Method of analysis of the sleeve dipole.

sponds to  $a_1 = a_2/2$ . Physically the condition  $\Omega_1 = \Omega_2 = 10$  has a meaning for the asymmetrical antenna with  $h_2 = 2h_1$ ;  $a_2 = 2a_1$ , but not for the antenna symmetrically driven by two equal generators. Thus, each of the two asymmetrical antennas in Fig. 9(b) may be analyzed, but the superposition of their solutions does not correspond *exactly* to any physically realizable structure since the outer thirds of each antenna cannot simultaneously have different radii. It is clear that the antenna of Fig. 9(d) also is approximated by a superposition of the two antennas in Fig. 9(b) except for the added complication of the shoulders at the driving points, of which no account is taken in the theory. In the following, the currents in the two antennas in Fig. 9(b) will be superimposed with the understanding that the results apply approximately to both Figs. 9(a) and 9(c), if a correction is made to take account of the shoulders in the latter.

The currents  $I_o$  and  $I_A$  in the central asymmetrical antenna in Fig. 8 with  $\Omega_1 = \Omega_2 = 10$  are obtained directly from Fig. 5. They are

$$\begin{aligned}\frac{I_o}{V_o^e} &= 3.98/51^\circ \\ &= 2.51 + j3.1 \text{ milliamperes per volt}\end{aligned}\quad (54a)$$

$$\begin{aligned}\frac{I_A}{V_o^e} &= 4.4/-81^\circ \\ &= 0.68 - j4.34 \text{ milliamperes per volt.}\end{aligned}\quad (54b)$$

Substitution in (53b) gives the following driving-point admittance at the terminals of each slice generator in the symmetrical two-generator antenna or at the terminals of the single-slice generator of half of the symmetrical structure erected vertically on an infinite conducting plane:

$$Y_{in} = (I_o + I_A)/V_o^e = (3.18 - j1.24) \times 10^{-3} \text{ mhos.} \quad (55a)$$

The corresponding input impedance is

$$Z_{in} = 273 + j106 \text{ ohms.} \quad (55b)$$

The distribution of current on the symmetrical two-generator antenna is obtained by superposition. It is

$$[I(z)]_{\text{symmetrical antenna}} = [I(z) + I(-z)]_{\text{asymmetrical antenna}} \quad (56)$$

since the current in the lower part of the middle unit in Fig. 8 is equal to the current in the upper part in the right-hand unit. The magnitude and phase of the current in the symmetrical two-generator antenna are given by

$$I = \sqrt{(I_t \cos \theta_t + I_b \cos \theta_b)^2 + (I_t \sin \theta_t + I_b \sin \theta_b)^2} \quad (57a)$$

$$\theta_I = \tan^{-1} [(I_t \sin \theta_t + I_b \sin \theta_b) / (I_t \cos \theta_t + I_b \cos \theta_b)] \quad (57b)$$

where

$$I_t = I_e e^{j\theta_t} \equiv I(z); \quad I_b = I_e e^{j\theta_b} \equiv I(-z). \quad (58)$$

The magnitude  $I$  and the phase-angle  $\theta_I$  referred to  $V_o^e$  are plotted in Fig. 10 for  $\beta_o h_1 = \pi/2$ ,  $\beta_o h_2 = \pi$ , and  $\Omega_1 \doteq \Omega_2 \doteq 10$ . The current in the central part of the antenna between the two generators is seen to be relatively great and to differ in phase from the currents in the outer parts by  $25^\circ$  on the average.

The admittance and impedance of a base-driven an-

tenna of electrical length  $\beta_o h = 3\pi/4$ ;  $\Omega = 10$ , base-driven over a perfectly conducting half space are

$$Y_o = (2.8 - j1.1) \times 10^{-3} \text{ mhos}; \quad Z_o = 310 + j122 \text{ ohms.} \quad (59)$$

These do not differ greatly from values (55a) and (55b) for the same antenna when driven at a height  $\lambda_o/8$  from the conducting plane.

The degree in which the impedance (55b) and the current in the upper half of Fig. 10 approximate the apparent terminal impedance of and the current on the sleeve dipole in Fig. 9(d) depends upon the size and the nature of the junction region where the coaxial line ends and the driving voltage is applied to the antenna. Usually, more-or-less extensive surfaces exist in the form of flat or curved surfaces with or without sharp edges on which charge can accumulate. Since these surfaces and transmission-line end effects are not considered in the analysis of the cylindrical antenna, it is to be expected that the measured apparent impedance of a sleeve dipole will differ considerably from the theoretical impedance. The additional chargeable surfaces are equivalent to a lumped capacitance in parallel with the theoretical impedance. The effect of a positive capacitance in parallel with a center-driven antenna is to decrease the magnitude of resistance and reactance and shift the curves toward shorter lengths. In particular, the positive reactance is reduced relatively much more than the negative reactance.

For example, suppose a susceptance  $B = \omega C = 3.2 \times 10^{-3}$  mhos is connected in parallel with a sleeve dipole for which the theoretical admittance and impedance are given by (55a) and (55b). The apparent admittance is

$$Y_{ina} = (3.18 + j1.96) \times 10^{-3} \text{ mhos.} \quad (60a)$$

The apparent impedance is

$$Z_{ina} = 228 - j141 \text{ ohms.} \quad (60b)$$

The broad-band properties of the sleeve dipole can not be investigated in general without extensive numerical computations of the input impedance for a variety of sleeve lengths and antenna lengths. However, the behavior as a function of frequency of the input impedance of the particular antenna for which the currents in Fig. 10 were determined may be carried out using available current and impedance data.

Suppose the frequency for which  $\beta_o h_1 = \pi/2$ ,  $\beta_o h_2 = \pi$ ,  $\beta_o l = \pi/4$  is  $f_o$ , and the associated angular velocity is  $\omega_o$ . For a variation in frequency from  $\omega_o - \Delta\omega$  to  $\omega_o + \Delta\omega$ , where  $\Delta\omega = \omega_o/4$ , the electrical lengths have the following ranges:

$$\frac{3}{4} \omega_o \leq \omega \leq \frac{5}{4} \omega_o: \quad \left\{ \begin{array}{l} \frac{3\pi}{8} \leq \beta_o h_1 \leq \frac{5\pi}{8} \\ \frac{3\pi}{4} \leq \beta_o h_2 \leq \frac{5\pi}{4} \\ \frac{3\pi}{16} \leq \beta_o l \leq \frac{5\pi}{16} \end{array} \right\}. \quad (61)$$

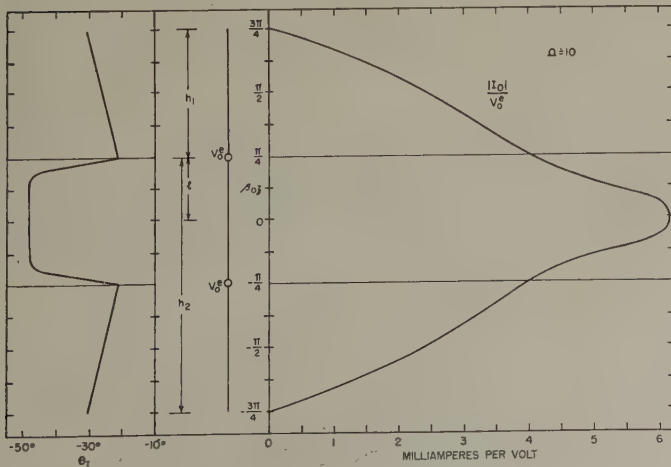


Fig. 10—Current distribution on the sleeve dipole.

The input impedance and admittance seen by the generator at  $z=l$  are given by (53a) and (53b). In order to determine the admittance or impedance over this range, it is necessary to know the currents  $I_o$  at the driving point  $z=l$  where the impedance is to be evaluated and  $I_A$  at  $z=-l$  in the asymmetrical antenna. These currents are available<sup>5</sup> for  $\beta_o h = \pi/2, 3\pi/4, \pi$ , and  $5\pi/4$ . It follows that in the range (61) the required currents may be determined for both extremes,  $\beta_o h_2 = 3\pi/4$  and  $5\pi/4$ , and for the middle value  $\beta_o h_2 = \pi$ . The current in the lower part 2 of the asymmetrical antenna is obtained from the current in the lower half of a symmetrical, center-driven antenna by multiplying by  $Z_2/(Z_1 + Z_2)$  where  $Z_1$  is the impedance of a symmetrical center-driven antenna of half length  $h_1$  and  $Z_2$  is the impedance of a symmetrical center-driven antenna of half length  $h_2$ . These values of impedance, as well as  $I_o/V_o^e = Y_2$  for the symmetrical antenna, may be determined over the entire range of variation, namely,  $3\pi/8 \leq \beta_o h_1 \leq 5\pi/8$ ;  $3\pi/4 \leq \beta_o h_2 \leq 5\pi/4$  using the second-order impedance and admittance curves.<sup>2</sup> The current  $I_A/V_o^e$  can be obtained from the second-order distribution curves<sup>5</sup> for  $\beta_o h = 3\pi/4, \pi$ , and  $5\pi/4$ . A reasonable estimate of the variation of  $I_A/V_o^e$  is obtainable by plotting curves of the magnitude and phase and of the real and imaginary parts of  $I_A$  as functions of  $\beta_o h$ . Using interpolated values from these curves the approximate input impedance  $Z_{in}$  of the sleeve dipole may be determined over the specified frequency range using the formula

$$\frac{1}{Z_{in}} = Y_{in} = 2 \left( \frac{I_o + I_A}{V_o^e} \right) \left( \frac{Z_2}{Z_1 + Z_2} \right). \quad (62)$$

Note that  $I_o$  and  $I_A$  in (62) are the values obtained in terms of a single generator with voltage  $V_o^e$ . Since the generator voltage at  $z=l$  in the sleeve dipole is  $V_o^e$ , the limiting case with  $l=0$  gives a generator of voltage  $2V_o^e$  at the center. Therefore, the factor 2 in (62) is required.

Curves showing the input impedance  $Z_{in} = R_{in} + jX_{in}$  are in Fig. 11 as a function of  $\beta_o h_2$  and as a function of  $\omega/\omega_o$ . For purposes of comparison the variation in  $Z_1 = R_1 + jX_1$  and  $Z_2 = R_2 + jX_2$  for symmetrical center-driven antennas of electrical half lengths in the ranges  $3\pi/8 \leq \beta_o h_1 \leq 5\pi/8$  and  $3\pi/4 \leq \beta_o h_2 \leq 5\pi/4$  are shown in Fig. 12. It is seen that the reactance of the sleeve dipole is remarkably constant near 110 ohms over the greater part of the range whereas  $X_1$  and  $X_2$  for the center-driven dipoles vary widely over positive and negative values.  $R_{in}$  for the sleeve dipole varies within 50 per cent of a mean value near 200 ohms whereas  $R_2$  ranges between 80 and 850 ohms, and  $R_1$  varies over a range of

nearly 100 per cent from a mean value of about 110 ohms. Clearly, the frequency response of the sleeve dipole is very much better from the point of view of

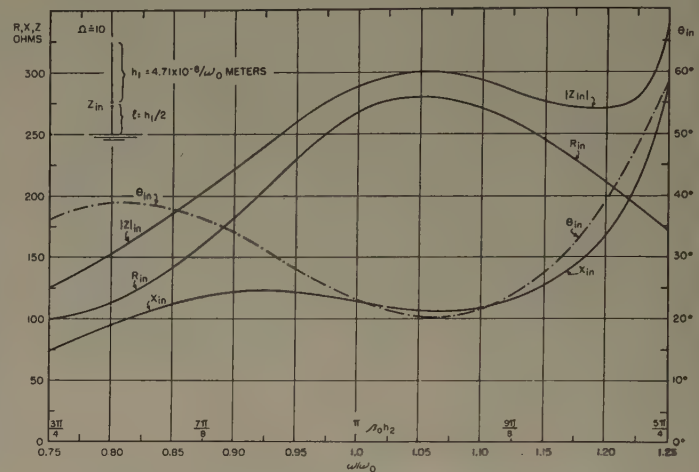


Fig. 11—Impedance behavior of the sleeve dipole.

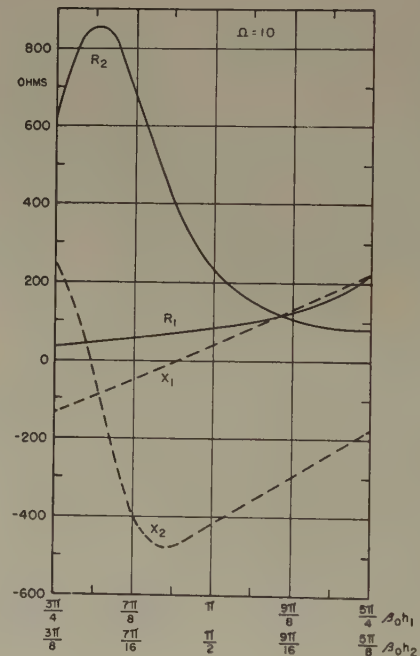


Fig. 12—Impedance behavior of the symmetrical center-driven antenna.

broad-band operation than the response of either a half-wave or a full-wave dipole. With an appropriately designed reactive matching network the standing-wave ratio on a line terminated in the sleeve dipole could be made to remain reasonably constant and small over a wide frequency range.

# Application of Correlation Analysis to the Detection of Periodic Signals in Noise\*

Y. W. LEE†, T. P. CHEATHAM, JR.†, AND J. B. WIESNER†, SENIOR MEMBER, IRE

**Summary**—The first part of this paper is a resumé of correlation analysis. An interpretation of the correlation function of a continuous random process is given in terms of the sum of the products obtained by representing the process as a discrete time series. An electronic correlator capable of performing the necessary mathematical operations to obtain correlation functions is described. The theory of detection of a periodic signal in random noise by the correlation method and a description of the application of the correlator to the theory are presented.

**P**RACTICAL APPLICATION of the recent developments in communication theory based upon the statistical concept of information has necessitated the study of characteristic functions of actual messages and noise as random processes and the development of techniques for their experimental determination. A statistical characteristic of messages or noise which has shown considerable effectiveness as a tool in the analysis of some communication problems is the correlation function.

As an application of the statistical approach to communication problems this paper presents the theory of detection of periodic signals in random noise by the method of correlation. The theory is supported by experimental results which have been obtained from an electronic correlator. The principles of the correlator are briefly presented. In order to provide sufficient background material, the first part of the paper contains a resumé of correlation analysis.

## I. CORRELATION ANALYSIS

### A. Autocorrelation of Random Processes

For a function belonging to a stationary random process<sup>1,2</sup> such as, for example, the fluctuations of shot noise or thermal noise with time as the independent variable, a characteristic known as the autocorrelation function has some interesting and important properties. If  $f_1(t)$  represents a member function of a stationary random process the autocorrelation function is defined as

$$\phi_{11}(\tau) = \lim_{T \rightarrow \infty} \frac{1}{2T} \int_{-T}^T f_1(t)f_1(t + \tau)dt. \quad (1)$$

\* Decimal classification: R361.211. Original manuscript received by the Institute, December 19, 1949; revised manuscript received, July 7, 1950. Presented, 1949 IRE National Convention, New York, N. Y., March 10, 1949.

This work has been supported in part by the Signal Corps, the Air Materiel Command, and ONR.

† Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge, Mass.

<sup>1</sup> M. C. Wang and G. E. Uhlenbeck, "On the theory of the Brownian motion," *Rev. Mod. Phys.*, vol. 17, pp. 323-342; April-July, 1945.

<sup>2</sup> H. M. James, N. B. Nichols, and R. S. Phillips, "Theory of Servomechanisms," McGraw-Hill Book Co., New York, N. Y., chaps. 6-8; 1947.

The function  $f_1(t)$ , which shall be called a random function, may or may not have a hidden periodic component. However, when it represents a shot noise or thermal noise, such a component does not exist. Although correlation functions may be defined for discrete random series, the functions concerned in this paper are continuous so that attention is confined to continuous random functions only. Because of the fact that expression (1) requires an averaging process with time as the independent variable, it is called a *time average* for  $\phi_{11}(\tau)$ .

It is known that an autocorrelation function is a continuous, even function with a maximum value at the origin which no other value of the function may exceed in magnitude. In the absence of a hidden periodic component, the function is asymptotic to the square of the mean of the random function. Consequently, if the mean is zero, the autocorrelation function tends to zero as the displacement  $\tau$  tends to infinity. Based upon these general properties a representative autocorrelation curve for a random phenomenon without hidden periodicity is roughly sketched in Fig. 1. If the random function is a voltage or current and a pure resistive load is assumed, then  $\phi_{11}(0)$  represents the total mean power,  $\phi_{11}(\infty)$  the dc power, and the difference, the power due to the fluctuating or random portion of the function.

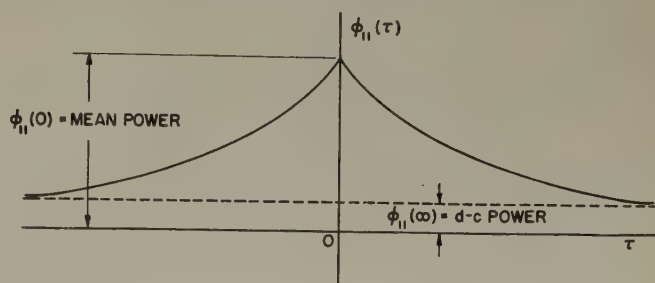


Fig. 1—Sketch of an autocorrelation function.

### B. Crosscorrelation of Random Processes

For two member functions  $f_1(t)$  and  $f_2(t)$  belonging to different stationary random processes but which are in some manner related to each other, a crosscorrelation function is defined by the expression

$$\phi_{12}(\tau) = \lim_{T \rightarrow \infty} \frac{1}{2T} \int_{-T}^T f_1(t)f_2(t + \tau)dt. \quad (2)$$

If the displacement  $\tau$  is given to  $f_1(t)$  instead of  $f_2(t)$ , then

$$\phi_{21}(\tau) = \lim_{T \rightarrow \infty} \frac{1}{2T} \int_{-T}^T f_2(t)f_1(t + \tau)dt. \quad (3)$$

Crosscorrelation is a measure of coherence between two random functions. For two random functions which are independently generated, crosscorrelation produces a constant which is the product of the individual mean values of the functions. Under this condition the functions are said to be incoherent; and in the particular case of either one having a zero mean, the crosscorrelation function is zero everywhere.

### C. Correlation of Periodic Functions

Correlation need not be confined to random phenomena. When the phenomenon under consideration is purely periodic, its autocorrelation function is defined by the same expression (1) in which  $f_1(t)$  is now the periodic function. Thus if  $f_1(t)$  has the general periodic form

$$f_1(t) = \frac{a_0}{2} + \sum_{n=1}^{\infty} a_n \cos(n\omega t + \theta_n), \quad (4)$$

the autocorrelation function is, by application of (1),

$$\phi_{11}(\tau) = \frac{a_0^2}{4} + \frac{1}{2} \sum_{n=1}^{\infty} a_n^2 \cos n\omega\tau. \quad (5)$$

This result brings out the interesting fact that the autocorrelation function of a periodic function is periodic, retains the fundamental and the harmonics, but drops all phase angles.

The crosscorrelation of periodic functions is defined by (2) and (3) in which  $f_1(t)$  and  $f_2(t)$  now represent the periodic functions. The crosscorrelation function of two periodic functions of the same fundamental frequency retains the fundamental and only those harmonics which are present in both, together with their phase differences.

### D. Ensemble Averages and the Ergodic Hypothesis

Instead of taking the time average of a member function of a stationary random process for the correlation function as it is done up to this point, the same result may be obtained by considering the process as a whole. The process consists of an infinite ensemble of functions which are generated by mechanisms of similar character. In this ensemble, let  $y_1$  be the random variable representing the values of the member functions observed at an arbitrary time  $t_1$ ,  $y_2$  those observed at time  $\tau$  later, and  $P(y_1, y_2; \tau)dy_1dy_2$  the joint probability that  $y_1$  lies in the range  $(y_1, y_1+dy_1)$  and  $y_2$  in the range  $(y_2, y_2+dy_2)$ . Because the process is stationary, the joint probability distribution is independent of  $t_1$ . A characteristic of the ensemble of particular interest is the mean product of the variables  $y_1$  and  $y_2$  as a function of the displacement  $\tau$ . This function is expressed as

$$\int_{-\infty}^{\infty} \int_{-\infty}^{\infty} y_1 y_2 P(y_1, y_2; \tau) dy_1 dy_2. \quad (6)$$

An important hypothesis in the theory of stationary random processes, known as the ergodic hypothesis,

states that the time average (1) is equivalent to the ensemble average (6). This hypothesis makes it possible to write

$$\phi_{11}(\tau) = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} y_1 y_2 P(y_1, y_2; \tau) dy_1 dy_2 \quad (7)$$

as the *ensemble average* for the autocorrelation function. Crosscorrelation functions obtained from ensemble consideration are expressed in a similar manner.

### E. Wiener's Theorem

The central theorem in the harmonic analysis of stationary random processes is Wiener's theorem<sup>3</sup> for autocorrelation. The theorem asserts that the autocorrelation function  $\phi_{11}(\tau)$  and the power density spectrum  $\Phi_{11}(\omega)$  of the random process are determinable one from the other by the integrals

$$\phi_{11}(\tau) = \int_{-\infty}^{\infty} \Phi_{11}(\omega) \cos \omega\tau d\omega \quad (8)$$

and

$$\Phi_{11}(\omega) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \phi_{11}(\tau) \cos \omega\tau d\tau. \quad (9)$$

In other words  $\phi_{11}(\tau)$  and  $\Phi_{11}(\omega)$  are Fourier cosine transforms of each other.

## II. AN ELECTRONIC METHOD FOR COMPUTING CORRELATION FUNCTIONS

### A. Theory

According to the definition of  $\phi_{11}(\tau)$  given in (1) a graphical method for its computation would consist of the following steps: (1) to displace  $f_1(t)$  by a small interval  $\tau_1$  resulting in  $f_1(t+\tau_1)$  as shown in Fig. 2; (2) to multiply these two functions continuously; (3) to inte-

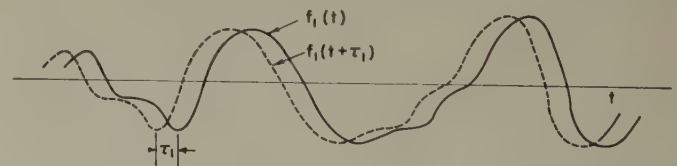


Fig. 2—Graphical method of obtaining an autocorrelation function.

grate the product over a long duration; and (4) to take the average value of the integral over the duration thus obtaining the points  $\phi_{11}(\tau_1)$  on the autocorrelation curve as illustrated in Fig. 3. Repetition of this procedure for

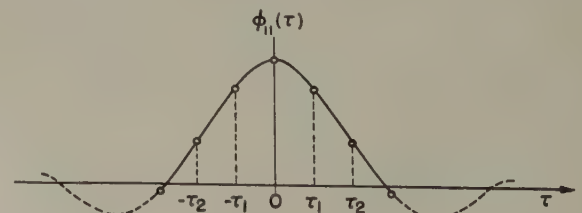


Fig. 3—Autocorrelation function obtained by graphical method.

<sup>3</sup> N. Wiener, "Generalized harmonic analysis," *Acta. Math.*, vol. 55, pp. 117-258; 1930.

different values of  $\tau$  determines as many points on the curve as necessary. To carry out this interpretation of autocorrelation by an electronic method would require means for delaying and storing the random function of a long duration. A satisfactory electronic device for continuous multiplication is also necessary. Unfortunately these essential requirements are difficult to meet at the present stage of development of electronic techniques.

However, a simpler electronic method which avoids these difficulties may be developed through the ensemble interpretation of the autocorrelation function. For the explanation of the method consider the random function  $f_1(t)$  in Fig. 4, which, for the time being, is as-

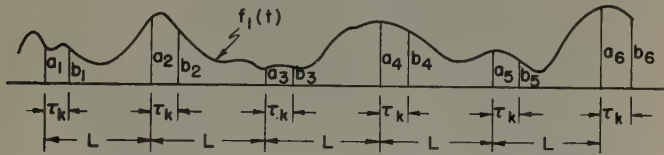


Fig. 4—Portion of a random function showing graphical representation of (12).

sumed to have no periodic component. Let the function be divided into sections each of duration  $L$ , and let it be assumed that the duration is so long that the set of values  $a_1, a_2, a_3, \dots$  of the function at the junction points are independent of one another. Under this condition the sections of the function may be considered as belonging to the member functions of a stationary random process. Now if  $b_1, b_2, b_3, \dots$  are the values of the sections at time  $\tau_k$  following the correspondingly numbered values in the first set, then the mean of the product of the two sets of values for  $n$  sections is

$$\frac{1}{n} \sum_{j=1}^n a_j b_j. \quad (10)$$

As the number of sections increases to infinity, the average quantity (10) tends to the value of the autocorrelation function of the random process for  $\tau = \tau_k$ . This interpretation of autocorrelation in the case of a physical situation is in agreement with the meaning of the ensemble average (7). Hence it is permissible to write

$$\phi_{11}(\tau_k) = \lim_{n \rightarrow \infty} \frac{1}{n} \sum_{j=1}^n a_j b_j. \quad (11)$$

In actual observation the number  $n$  can be made large, but it remains finite so that the autocorrelation curve at  $\tau = \tau_k$  has the approximate value

$$\phi_{11}(\tau_k) \cong \frac{1}{n} \sum_{j=1}^n a_j b_j. \quad (12)$$

Clearly the error in approximation depends upon the number  $n$ . The calculation of this error is discussed in a later section when correlation analysis is applied to the problem of signal detection.

If a former restriction on  $f_1(t)$  is removed so that it now has an additive periodic component but the sections still have a fixed duration  $L$ , then the values

$a_1, a_2, a_3, \dots$  are no longer independent of one another. However, from a practical point of view, if the ratio of the period of the periodic component wave and the duration  $L$  is not such that the junction points of the sections always occur at only a few fixed locations on each period of the wave, the expression (12) should hold for a random function with a periodic component.

When the duration  $L$  is so short that the values  $a_1, a_2, a_3, \dots$  are in some manner related to one another even in the absence of a periodic component, the formula (12) may be shown to remain valid on the basis of time averaging for the correlation function carried out in accordance with the principles of sampling. The remarks just made in regard to the presence of periodicity applies in the present case.

Keeping the section duration  $L$  a constant is a great convenience in actual computation by electronic means. But if it were made a random variable the approximate expression (12) may still be applied in the point-by-point determination of the correlation curve.

Although the expression (12) is applicable in the several different situations discussed, it does not necessarily give the same value in all cases for the same number  $n$ . The calculation of error in each case depends upon the manner in which the values  $a_j$  and  $b_j$  are obtained.

### B. Electronic Techniques

The electronic method embodying the principles discussed for the computation of correlation functions is indicated in the diagram of Fig. 5 and a brief description follows. The irregular continuous wave form (a) repre-

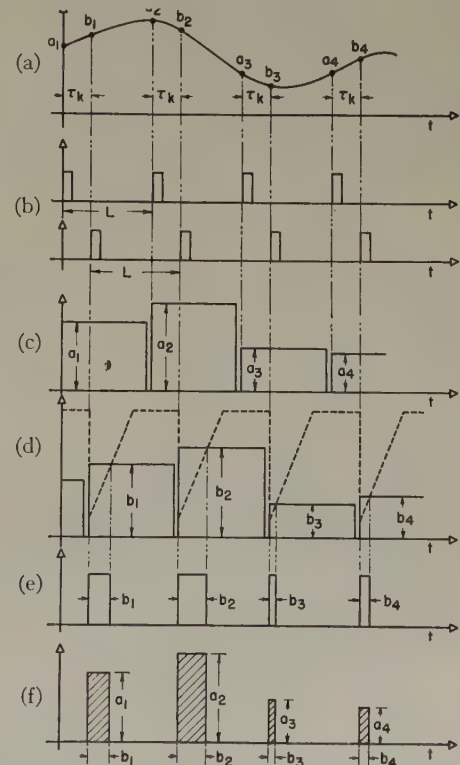


Fig. 5—Characteristic wave forms of electronic correlator.

sents a portion of the random function for which the autocorrelation curve is to be computed. Two sets of periodic timing pulses (b) both with the period  $L$ , are derived from a sine-wave master oscillator. The second set of timing pulses is displaced by a delay network from the first set by the adjustable duration  $\tau_k$ . By means of the first set of timing pulses, the values  $a_1, a_2, a_3 \dots$  are obtained. A boxcar wave form (c) is then generated in such a way that its discretely varying amplitudes are proportional to these values. Similarly another boxcar wave form (d) is formed from the set  $b_1, b_2, b_3 \dots$ . A sawtooth sampling wave converts the amplitudes of the second boxcar wave into a series of duration-modulated pulses (e). Clearly these pulses represent the  $b$  series. By means of a gating circuit the wave forms (c) and (e) are multiplied giving wave form (f) which is a series of pulses whose amplitudes represent the  $a$  series and durations the  $b$  series. The integration of wave form (f) by an integrating circuit for a large number of samples  $n$  results in a value which is proportional to the value  $\phi_{11}(\tau_k)$  as expressed by (12).

An electronic correlator<sup>4</sup> based upon the theory and techniques discussed has been developed at the Research Laboratory of Electronics, Massachusetts Institute of Technology. A photograph of the correlator is shown in Fig. 6. The displacement  $\tau_k$  is automatically adjusted in discrete steps as the correlation curve is computed point by point each for the same number of samples. The results are presented graphically by a recording meter. The machine has been designed for crosscorrelation as well as autocorrelation.

### III. DETECTION OF SINUSOIDAL SIGNAL IN RANDOM NOISE

#### A. Principles of Detection by Autocorrelation

Consider an additive mixture of a sinusoidal signal  $S(t)$  and a random noise  $N(t)$ . Let the mixture be written as

$$f_1(t) = S(t) + N(t). \quad (13)$$

By application of the defining equation (1) the autocorrelation function of  $f_1(t)$  is

$$\begin{aligned} \phi_{11}(\tau) &= \lim_{T \rightarrow \infty} \frac{1}{2T} \int_{-T}^T [S(t) + N(t)][S(t+\tau) + N(t+\tau)] dt \\ &= \phi_{SS}(\tau) + \phi_{NN}(\tau) + \phi_{SN}(\tau) + \phi_{NS}(\tau). \end{aligned} \quad (14)$$

The first two terms of the result are the autocorrelation functions of the sinusoid and the noise, respectively, and the last two terms are their crosscorrelation functions. As a matter of convenience, let the means of both  $S(t)$  and  $N(t)$  be zero. If  $S(t)$  has the form

$$S(t) = E_m \cos(\omega t + \theta), \quad (15)$$

<sup>4</sup> Further details are given in Technical Report No. 122, T. P. Cheatham, "Experimental Determination of Correlation Functions and Their Application in the Statistical Theory of Communications," Research Laboratory of Electronics, MIT (to be published).

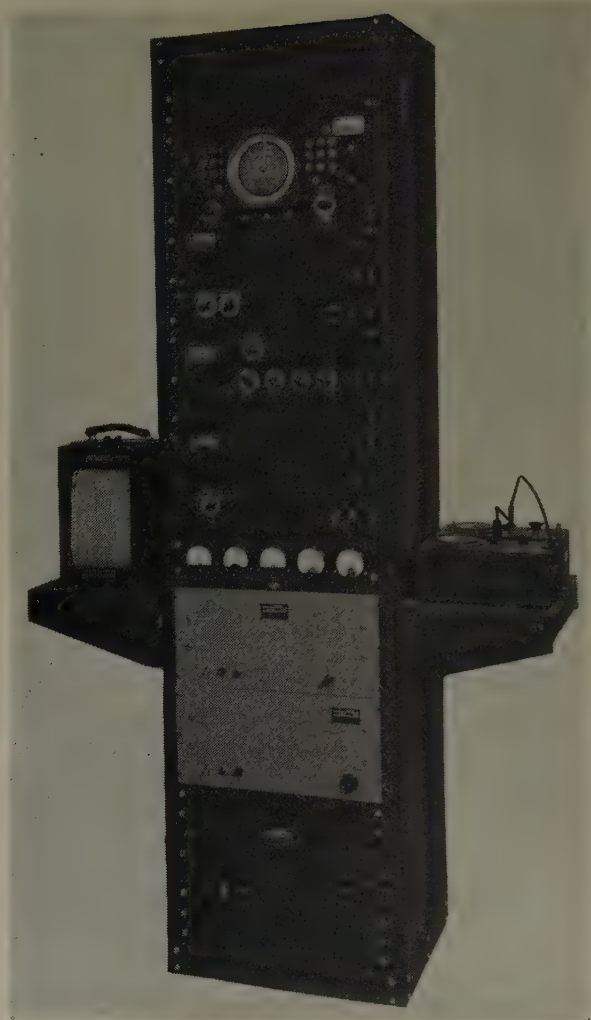


Fig. 6—MIT electronic correlator.

then according to the general expression (5)

$$\phi_{SS}(\tau) = E^2 \cos \omega \tau \quad (16)$$

where  $E$  is the rms value of the sinusoid. The function  $\phi_{NN}(\tau)$  is nonperiodic and tends to zero as  $\tau$  tends to infinity. Because of incoherence between signal and noise the crosscorrelation terms vanish in view of a previous discussion. From these considerations the general appearance of the autocorrelation function (14) may be sketched as in Fig. 7.

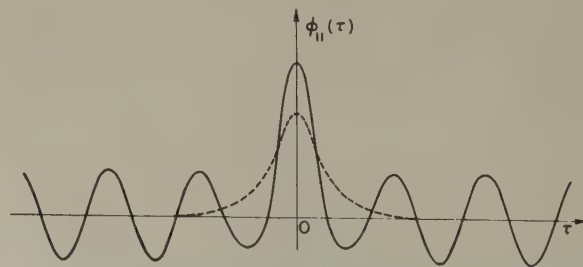


Fig. 7—Autocorrelation function of sine wave plus random noise. Dotted curve is component due to random noise.

The fact that the autocorrelation of random noise approaches zero as the displacement becomes large

whereas that of a periodic function persists as a periodic function, suggests that in the region of  $\tau$  where noise autocorrelation is essentially zero, the presence of a periodic wave in the input of the correlator should be evidenced by a periodic variation in its output. Theoretical considerations show that a periodic signal masked by random noise can be detected by autocorrelation however small the signal may be. However, the theoretical possibility requires, among other things, an infinite time for observation.

### B. Nature of Correlator Output

In view of the fact that the correlator operates according to sampling theory and in view of the finite time of observation in a practical situation, it is important to determine to what extent a periodic signal may be detected by this method. When the input of the correlator is a sine wave plus random noise its output graph for displacement  $\tau$  sufficiently beyond the coherent range of noise is as shown in Fig. 8. Let it be as-

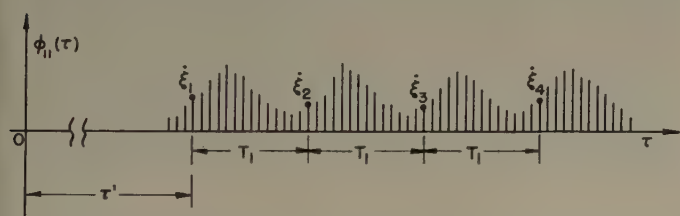


Fig. 8—Output graph of electronic correlator for large values of  $\tau$ .

sumed that there are sufficient number of points in each cycle of the output curve for the identification of a set of points  $\xi_1, \xi_2, \xi_3, \dots$ , as indicated, which are located at a corresponding position in each cycle. Other sets of points may be formed in a similar manner. Let each point be the average of  $n$  sample products. The random variable from which the sample mean  $\xi_1$  of size  $n$  is drawn is the variable

$$\begin{aligned} \xi &= [S(t) + N(t)][S(t + \tau') + N(t + \tau')] \\ &= S(t)S(t + \tau') + N(t)N(t + \tau') \\ &\quad + S(t)N(t + \tau') + N(t)S(t + \tau') \end{aligned} \quad (17)$$

where  $\tau'$  is the time displacement from the origin to the point  $\xi_1$ . The random variables from which the other sample means  $\xi_2, \xi_3, \dots$  are drawn are similarly written with  $\tau'$  increased by integral multiples of the period of the sinusoid. These random variables have the same distribution because of incoherence between  $N(t)$  and  $N(t + \tau)$  for  $\tau > \tau'$  and of the particular values of successive displacements.

To simplify the discussion, assume for the moment that the sampling process in the correlator is random. Actually, of course, the samples are taken at periodic intervals. Under the temporary assumption, the sample means  $\xi_1, \xi_2, \xi_3, \dots$  are independent and are drawn from variables with the same distribution. It is known in statistical theory that in such circumstances the sample mean, as a random variable, has the mean of the vari-

able  $\xi$  of (17) and the variance that of  $\xi$  divided by the sample size  $n$ .

### C. Correlator Output Noise

In determining the gain in signal-to-noise ratio it is necessary to discuss the nature of the output noise. Whereas the input noise is a continuous fluctuation, the output "noise" consists of the dispersion of a set of discrete points about a mean. The variance of the random sample means such as the set  $\xi = \xi_1, \xi_2, \xi_3, \dots$  is a measure of the output noise. Clearly the input and output of the correlator cannot be compared on an instantaneous time basis as is possible in the case of a wave filter. The correlator output is a function of the displacement  $\tau$ , not the independent variable time, for each value of which an average product is taken over a long duration.

If  $E$  is the rms value of the sinusoidal signal and  $\sigma$  is the standard deviation of the random noise (variance  $= \sigma^2$ ), and if both signal and noise have a zero mean, it may be shown that the variance of the sample mean  $\xi$  as a random variable has the expression

$$\sigma_{\xi}^2 = \frac{1}{n} \left( \frac{1}{2} E^4 + 2E^2\sigma^2 + \sigma^4 \right). \quad (18)$$

This result has been obtained in accordance with the known properties of a sample mean already stated. One important property of the variance (18) is its independence of the displacement  $\tau$  in the region under consideration.

### D. Gain in Autocorrelation

Let the rms noise-to-signal ratio at the input be  $\rho_i$  so that

$$\rho_i = \frac{\sigma}{E}. \quad (19)$$

In decibel units, the input signal-to-noise ratio is

$$R_i = 20 \log_{10} \frac{1}{\rho_i} \text{ db}. \quad (20)$$

According to (15) and (16) the ideal output signal, as a function of  $\tau$ , has the rms value

$$S_o = \frac{E^2}{\sqrt{2}}. \quad (21)$$

The output noise  $N_o$  in rms value of the sample means about the ideal mean is the square root of the variance give in (18). Therefore

$$N_o = \sigma_{\xi} = \sqrt{\frac{1}{n} \left( \frac{1}{2} E^4 + 2E^2\sigma^2 + \sigma^4 \right)} \quad (22)$$

and the correlator output rms signal-to-noise ratio becomes

$$\frac{S_o}{N_o} = \frac{\sqrt{n} E^2}{\sqrt{E^4 + 4E^2\sigma^2 + 2\sigma^4}}. \quad (23)$$

In decibel units the ratio is

$$R_{oa} = 20 \log_{10} \frac{S_o}{N_o} \text{ db}$$

$$= 10 \log_{10} \frac{n}{1 + 4\rho_i^2 + 2\rho_i^4} \text{ db.} \quad (24)$$

A curve of this expression for  $n = 60,000$  is given in Fig. 9.

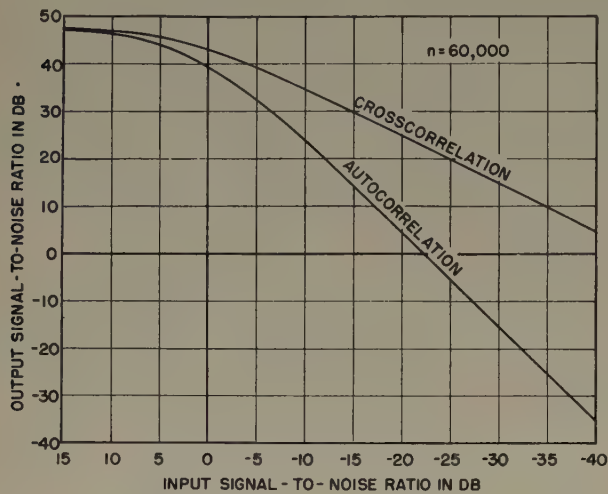


Fig. 9—Improvement in signal-to-noise ratio through autocorrelation and crosscorrelation for sinusoid in random noise.

### E. Periodic Sampling

Analysis shows that, except when the sampling period happens to be such that samples are selected at points corresponding to only a small fraction of a cycle of the sinusoid or at points restricted to a few locations in a cycle, periodic sampling reduces the variance (18) by decreasing its first term. This quantity originates from the purely periodic component, that is, the first term of the variable (17). For a large sample size in periodic sampling the first term of (18) is negligible. On the other hand, if it may be assumed that there is sufficient variation in the duration of switching time in the correlator for each change of  $\tau$ , so that the starting phase of the sinusoid for each point of the output curve is random, the second term of (18) remains the same as before. The last term of (18) is independent of the manner of sampling.

Under these conditions the correlator output signal-to-noise ratio is

$$R_{oa}' = 10 \log_{10} \frac{n}{2\rho_i^2(2 + \rho_i^2)} \text{ db.} \quad (25)$$

For large values of  $\rho_i$  there is practically no difference between random sampling and periodic sampling as far as the ratio is concerned.

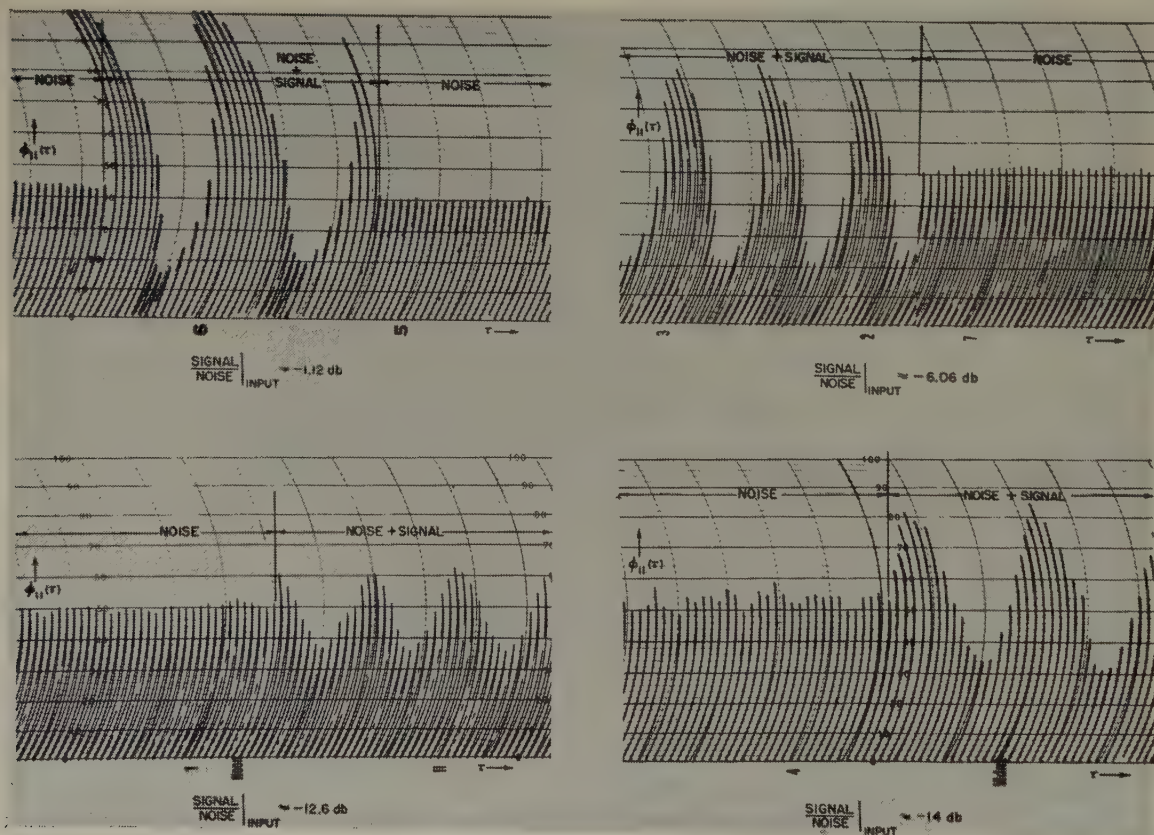


Fig. 10—Experimental results from electronic correlator showing gain in output signal-to-noise ratio for input of sinusoid plus random noise at several ratios.

### F. Experimental Results

To verify the analytical results in 8-kc sine wave plus random noise derived from a Type-884 gas tube was supplied to the electronic correlator. The correlator output curves as presented by the recorder for input signal-to-noise ratios of  $-1.12$  db,  $-6.06$  db,  $-12.6$  db, and  $-14$  db are shown in Fig. 10. The comparatively flat portions of the curves represent the autocorrelation of noise alone while the other portions indicate clearly the presence of a sine wave. In this experiment the sampling rate was 500 sample products per second and the sample size was 60,000.

### G. Gain in Crosscorrelation

If the frequency of the sinusoidal signal is known, advantage may be taken of this information for attaining a further gain in signal-to-noise ratio. Instead of autocorrelation, a local sinusoidal wave of the known frequency is generated for cross correlating with the incoming corrupted signal. The advantage of crosscorrelation over autocorrelation may be seen from an examination of the random variable from which the correlator output is obtained. For crosscorrelation the input function is

$$f_1(t) = S_1(t) + N_1(t) \quad (26)$$

and the local wave is

$$f_2(t) = S_2(t) \quad (27)$$

so that the variable from which samples are drawn is

$$\begin{aligned} \eta &= [S_1(t) + N_1(t)]S_2(t + \tau) \\ &= S_1(t)S_2(t + \tau) + N_1(t)S_2(t + \tau). \end{aligned} \quad (28)$$

Comparison of this variable and the variable in (17) shows that because of the reduction of the number of terms from four to two the noise in crosscorrelation should be much less. In fact, the principal contribution to the variance, which is the measure of output noise, is accounted for by the pure noise term  $N(t)N(t+\tau')$  in (17) so that its absence in (28) permits a substantial gain.

By calculations similar to those for autocorrelation, the output signal-to-noise ratio in crosscorrelation by random sampling may be shown to have the expression

$$R_{oc} = 10 \log_{10} \frac{n}{1 + 2\rho_i^2} \text{ db.} \quad (29)$$

A curve for this equation when  $n=60,000$  is given in Fig. 9.

A comparison of (24) and (29) shows clearly that for the same sample size, crosscorrelation has a gain over autocorrelation of the amount

$$G = 10 \log_{10} \frac{1 + 4\rho_i^2 + 2\rho_i^4}{1 + 2\rho_i^2} \text{ db.} \quad (30)$$

Fig. 11 contains a curve of this expression.

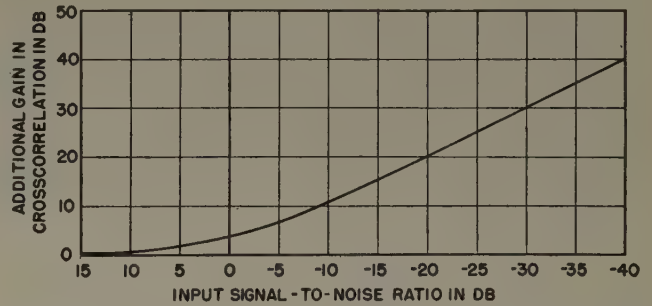


Fig. 11—Additional gain in signal-to-noise ratio through crosscorrelation as compared with autocorrelation (difference of the two curves in Fig. 9.)

Under conditions of Section III, Part E, periodic sampling reduces (29) to

$$R_{oc}' = 10 \log_{10} \frac{n}{2\rho_i^2} \text{ db} \quad (31)$$

and (30) to

$$G' = 10 \log_{10} (2 + \rho_i^2) \text{ db.} \quad (32)$$

### IV. REMARKS

The method and technique of detecting a periodic wave in random noise presented here may be regarded as a type of filtering in the time domain. It is well to emphasize that theoretically no claim can be made that the method of correlation is superior to conventional filtering in the frequency domain in the sense of accomplishing in the time domain what is impossible in the frequency domain. Such a claim would be false, since it has already been stated that the correlation function and power-density spectrum of a stationary random process are uniquely related by a Fourier transformation, and since it is well known that a similar relationship holds for a periodic phenomenon. Nevertheless, from an engineering point of view, many equivalent operations may be more practicable and feasible in the time domain. For instance, a "zero" bandwidth filter in the frequency domain corresponds to an extremely stable oscillator in the time domain. At the present time it is much easier to build a stable oscillator than to build a network having zero bandwidth.

### ACKNOWLEDGMENTS

The writers wish to express their appreciation for the valuable suggestions and assistance given by M. Cerrillo, E. R. Kretzmer, C. A. Stutt, and M. Loewenthal.

# The Traveling-Wave Cathode-Ray Tube\*

K. OWAKI†, S. TERAHATA†, T. HADA†, AND T. NAKAMURA†

**Summary**—We have constructed cathode-ray tubes utilizing traveling waves, whose deflection sensitivity can be made independent of frequency by equalizing the phase velocity of the wave and the electron velocity. According to the theoretical analysis, the sensitivity of this tube is increased to  $10^{-2}\text{mm/v}$  at 30 kMc without any difficulty. The results of the theoretical analysis are proved experimentally by the inversion spectrums.

As examples of its application, measurements of amplitude modulation degree and observations of uhf voltage wave form are described in this paper.

## INTRODUCTION

SEVERAL PAPERS have hitherto been published about cathode-ray tubes or oscillographs for measurement in the region of uhf. For instance, Hollmann<sup>1,2</sup> together with Von Ardenne, has studied the Lissajous figures in uhf region using a specially designed oscillograph, in which a very fine electron beam was focussed electrostatically and was deflected by a pair of extremely small deflecting plates. Recently, Lee<sup>3</sup> has succeeded in taking photographs of voltage waves at 10,000 Mc using an oscillograph of three elements with a magnetic lens, operated similarly to that of Hollmann.

These cathode-ray tubes or oscillographs adopt either electron beams of extremely high velocity or deflecting plates of very small size in order to render it possible to neglect the transit-time effect (transit-time effect of the first kind) under the deflecting plates.

On the other hand, in 1943 the authors of this paper proposed a new method of electron deflection based upon the principle of velocity modulation. In this cathode-ray tube, the variation in velocity of the electron beam, which is proportional to the instantaneous uhf voltage, is transformed into the transverse deflection of the beam.

Recently we have developed a traveling-wave cathode-ray tube which makes use of a pair of parallel wires folded repeatedly instead of the usual deflecting plates. The electron beam is passed between these wires and is deflected by the electric field between the parallel wires.

A similar idea is seen in Haeff's patent<sup>4</sup> and J. R. Pierce's paper,<sup>5</sup> in which the former utilizes distributed constant circuits and the latter lumped constant circuits for the realization of deflecting system.

## CONSTRUCTION

The construction of this tube is similar to that of conventional types, except for the deflecting device. As is shown in Fig. 1, two pairs of deflecting devices are constructed with parallel wires folded repeatedly, the respective fields of deflection being so arranged as to be at right angles to each other.

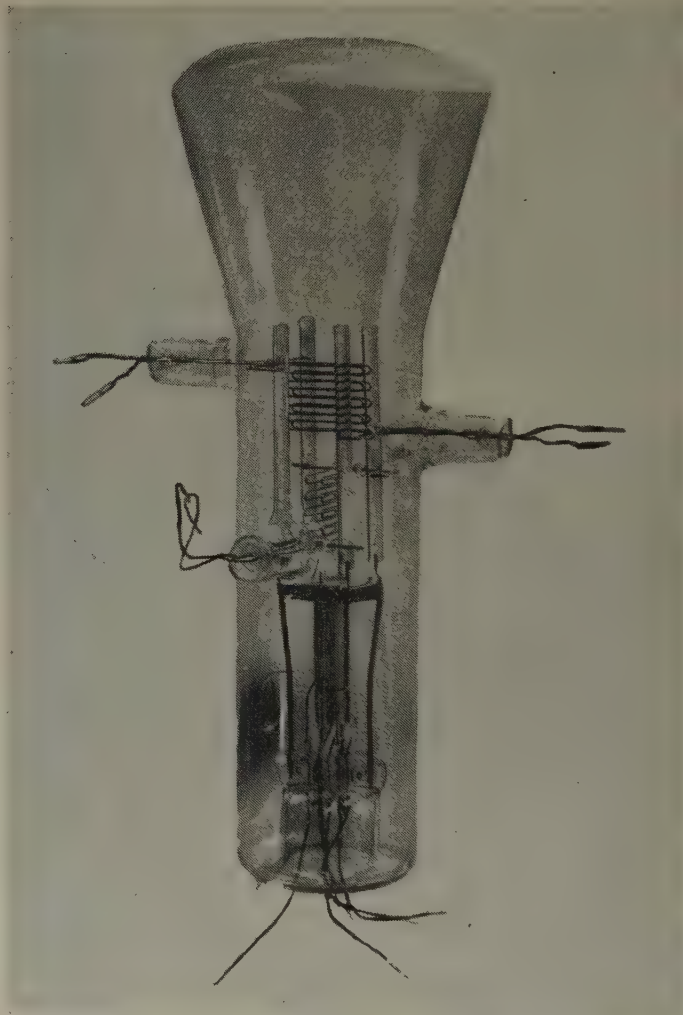


Fig. 1—Construction of the traveling-wave cathode-ray tube.

## PRINCIPLES

The electron travels along the axis of the deflecting device from *A* to *B* between the two parallel wires, as shown in Fig. 2. The electromagnetic wave,  $V_m \sin \omega t$ , is made to travel from *P* to *Q* along the parallel wires. Because of the folded nature of the transmission wires which constitute the deflecting device, the phase velocity *u* of the electromagnetic wave along the axis *AB* becomes smaller than that on the parallel wires,

\* Decimal classification: R138.31×R339.2. Original manuscript received by the Institute, November 30, 1949; revised manuscript received, July 6, 1950.

† Kobe Kogyo Corporation, Kobe, Japan.

<sup>1</sup> H. E. Hollmann, "Mikrowellen-Oszillographie," *Hoch. Frequenz. Tech.*, vol. 54, pp. 188-190; December, 1939.

<sup>2</sup> H. E. Hollmann, "Ultra-high-frequency oscillography," *Proc. I.R.E.*, vol. 28, pp. 213-219; May, 1940.

<sup>3</sup> G. M. Lee, "A three-beam oscillograph for recording at frequencies up to 10,000 megacycles," *Proc. I.R.E.*, vol. 34, pp. 121W-127W; March, 1946.

<sup>4</sup> A. Haeff, U. S. Patent No. 2,064,469.

<sup>5</sup> J. R. Pierce, "Traveling-wave oscilloscope," *Electronics*, vol. 22, p. 97; November, 1949.

and is given by

$$u = \frac{a}{l} \times c \quad (1)$$

where

$a$  = spacing between adjacent parallel wires

$l$  = folded length of the wire between adjacent points on the axis

(0-1; 1-2  $\cdots$   $n-1-n$ : refer to Fig. 2)

$c$  = light velocity.

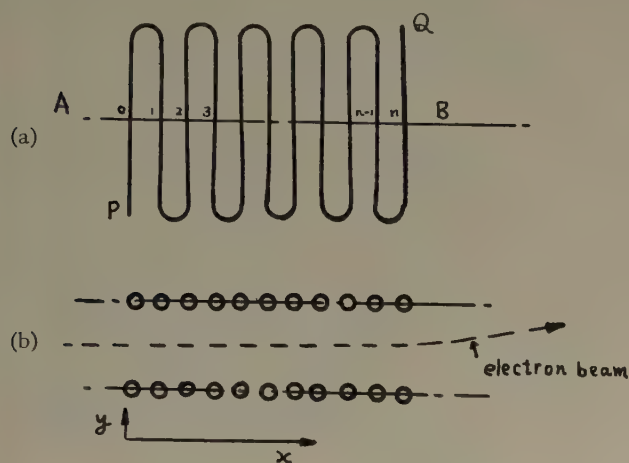


Fig. 2—Diagrammatical illustration of deflecting system. (a) View of the folded parallel transmission lines from above. (b) Side view.

Suppose that the velocity  $v_0$  of the electron along the axis ( $x$  axis) is constant, and let  $u = v_0$ . Under this condition, electrons passing through the deflecting device are subject to a constant accelerating field in direction, and magnitude, if the attenuation of the wave is neglected. This is because the transit time of the electron from A to B is equal to the propagating time of the electromagnetic wave.

We shall proceed to a mathematical analysis of the deflection of the electron. As is shown in Fig. 2, the  $x$ -axis is taken along the axis of the deflecting device, and the  $y$  axis is taken to coincide with the direction of the deflecting force which is perpendicular to the  $x$  axis. The electron, traveling with the velocity  $v_0$  along the  $x$  axis, is deflected in the direction of the  $y$  axis by the electric field due to the voltage on the deflecting device. If the deflecting device consists of a pair of plate electrodes, the electric field in the  $y$  direction at any point on the  $x$  axis will be a function of time only. But in this case, the electric field is a function of both  $t$  and  $x$ .

Here, we investigated the potential distribution along the deflecting device by using an electrolytic bath. First, the equipotential curves on a plane perpendicular to the parallel wires which contain the  $x$  axis of the device were obtained. Then, an approximate transverse field distribution along the  $x$  axis was deduced from the potential difference between two points having an infinitesimal distance separation on both sides of the  $x$  axis.

Fig. 3 shows an example of equipotential curves at a phase difference of  $40^\circ$  between the adjacent parallel wires, in which 18 circles represent the electrodes and the number in the circles denote their respective applied voltage referred to the maximum value of uhf voltage taken as 100. Fig. 4 (a), (b), (c), (d), and (e) show distributions of transverse potential gradients along the  $x$  axis obtained from the equipotential curves corresponding to phase differences of  $40^\circ$ ,  $60^\circ$ ,  $120^\circ$ ,  $180^\circ$ , and  $240^\circ$  between adjacent parallel wires, respectively. Full lines and dotted lines represent results of experiments and sinusoidal curves, respectively. These curves are almost identical with the exception of the end disturbances. However, in these results the potential variation along the parallel wires is left out of consideration, so that the results are applicable only when  $l$  is very small compared with the wavelength.

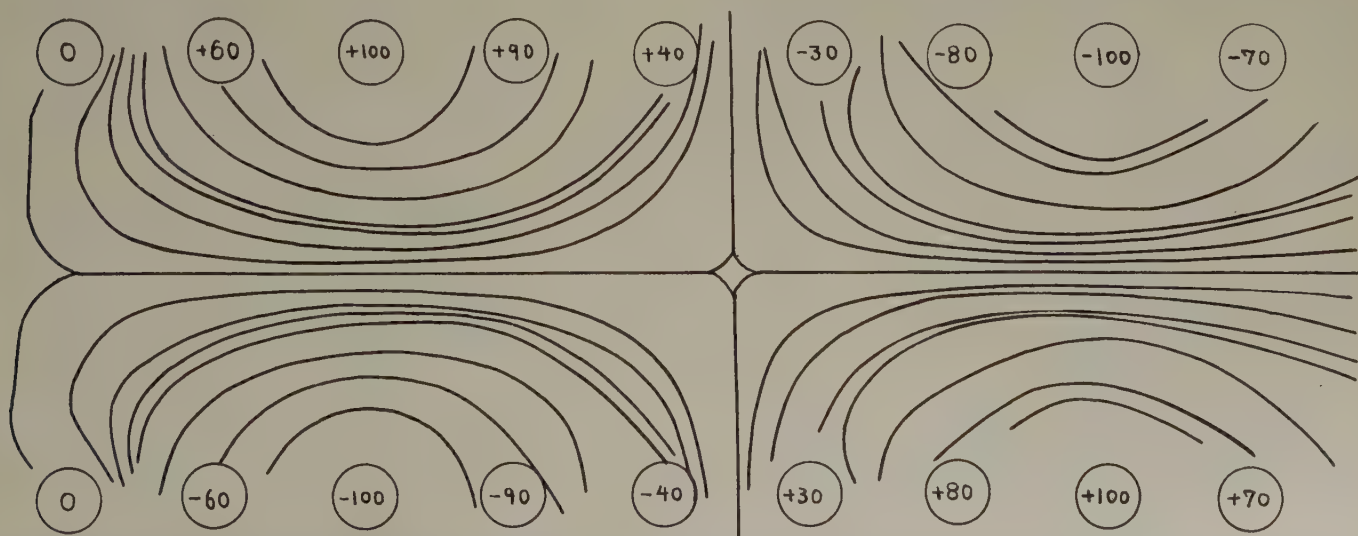


Fig. 3—Equipotential curves on the plane perpendicular to parallel wires which contain the  $x$  axis of the deflecting device. This figure corresponds to (b) of Fig. 2. The circles indicate the cross section of wires, and the numerical figure in these circles show the voltage of the wires for phase difference of 40 degrees between adjacent parallel wires.

Now, from the above results, it may safely be assumed that the voltage waves along the axis of the deflecting device is expressed by a cosinusoidal form. Also, the wave propagating along the parallel wires from  $P$  to  $Q$  will, in general, be partially reflected at  $Q$ ; hence at a point on the axis of the deflecting device the electric field will be composed of the incident and reflected waves, leading to the resultant potential difference between the parallel wires at a point  $x$  on the axis to be represented as follows:

$$V_{(x,t)} = V_m \cos \omega \left( t - \frac{x}{u} \right) + (1 - \beta) V_m \cos \left[ \omega \left\{ t - (2an - x) \frac{1}{u} \right\} - \phi \right] \quad (2)$$

where

$V_m$  = maximum value of uhf voltage

$u$  = velocity of propagation of the electromagnetic wave along the axis of the deflecting device as described above, i.e., phase velocity

$\beta, \phi$  = attenuation coefficient and the phase angle or the electromagnetic wave at  $B$  when reflection occurs at  $Q$ .

The transverse electric field  $E$  on the  $x$  axis may be obtained as follows:

$$E_{(x,t)} = \frac{k}{D} \left\{ V_m \cos \omega \left( t - \frac{x}{u} \right) + (1 - \beta) V_m \cos \left[ \omega \left( t - (2an - x) \frac{1}{u} \right) - \phi \right] \right\} \quad (3)$$

where

$D$  = distance between two parallel wires

$k$  = ratio of the electric field between the two parallel wires and that between two imaginary parallel plates extended infinitely.

The questions of motion of electron are expressed as follows:

$$\frac{d^2x}{dt^2} = 0 \quad (4)$$

$$\frac{d^2y}{dt^2} = \frac{e}{m} \frac{kV_m}{D} \left\{ \cos \omega \left( t - \frac{x}{u} \right) + (1 - \beta) \cos \left[ \omega \left\{ t - (2an - x) \frac{1}{u} \right\} - \phi \right] \right\} \quad (5)$$

From (4) we obtain

$$\frac{dx}{dt} = v_0(\text{const}). \quad (6)$$

Substituting  $x = (t - t_0)v_0$  in (5), and integrating

$$\frac{dy}{dt} = \frac{2e}{m} \frac{kV_m}{D} \frac{u}{\omega} \left[ \frac{1}{(u - v_0)} \sin \frac{\omega}{u} \left( 1 - \frac{v_0}{u} \right) (t - t_0) \cdot \cos \frac{\omega}{2} \left\{ t - \frac{v_0}{u} (t - t_0) + t_0 \right\} \right]$$

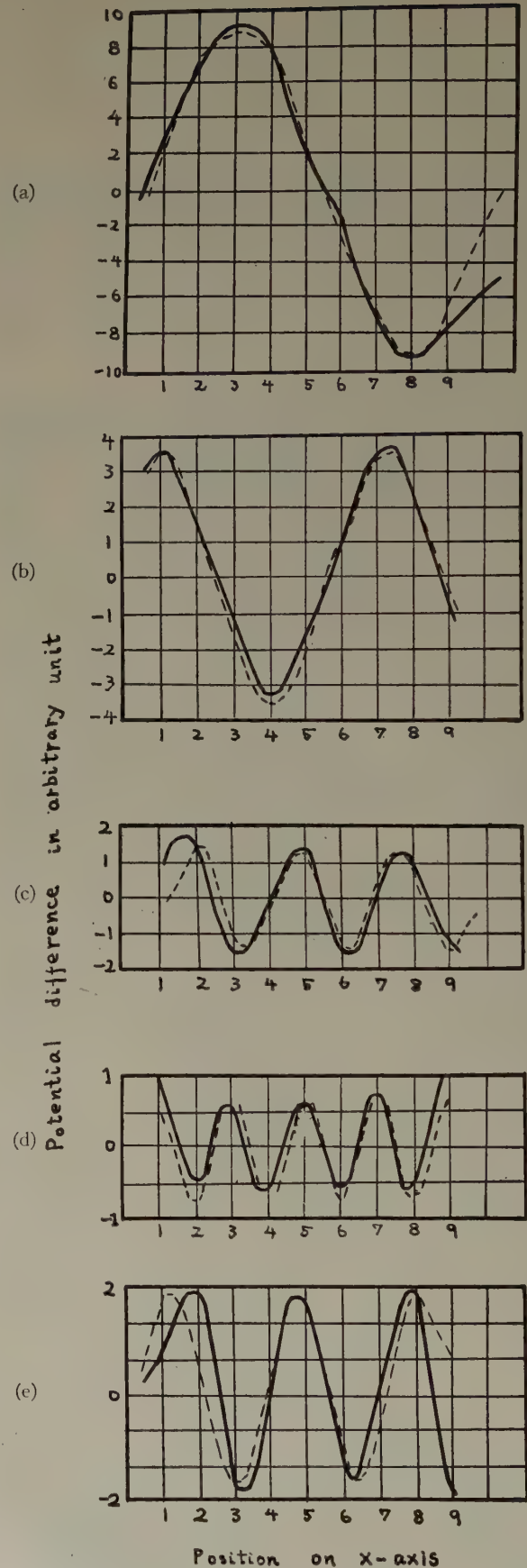


Fig. 4—Distribution of transversal potential gradient along the  $x$  axis for various phase difference of voltage between adjacent parallel wires. (a) Phase difference of adjacent parallel wires, 40 degrees. (b) 60 degrees. (c) 120 degrees. (d) 180 degrees. (e) 240 degrees.

$$\begin{aligned}
& + (1 - \beta) \left\{ \frac{1}{(u + v_0)} \sin \frac{\omega}{2} \left( 1 + \frac{v_0}{u} \right) (t - t_0) \right. \\
& \cdot \cos \frac{1}{2} \left[ \omega \left\{ \left( \frac{v_0}{u} - 1 \right) (t - t_0) \right. \right. \\
& \left. \left. + 2t - \frac{4an}{u} \right\} - 2\phi \right] \left. \right\} \quad (7)
\end{aligned}$$

where,  $t_0$  = time of departure of the electron at  $A$ . Substituting  $c = (t - t_0)v_0 = an$  in (7), the transverse velocity at the end of deflecting device can be obtained as follows,

$$\begin{aligned}
\frac{dy}{dt(x=an)} &= \frac{2e}{m} \frac{kV_m}{D} \frac{u}{\omega} \left[ \frac{1}{(u - v_0)} \sin \frac{an\omega(u - v_0)}{2uv_0} \right. \\
&\cdot \cos \left\{ \omega t - \frac{an\omega(u + v_0)}{2uv_0} \right\} \\
&+ \frac{(1 - \beta)}{(u + v_0)} \sin \frac{an(u + v_0)}{2uv_0} \\
&\cdot \cos \left\{ \omega t - \frac{an\omega(u + 3v_0)}{2uv_0} - \phi \right\} \left. \right]. \quad (8)
\end{aligned}$$

This equations can be rewritten thus:

$$\frac{dy}{dt(x=an)} = K \cdot \cos(\omega t - \theta), \quad (9)$$

where

$$\begin{aligned}
K &= \frac{2ekV_m}{mD\omega} \left\{ \frac{1}{(u - v_0)} \sin^2 \frac{an\omega(u - v_0)}{2uv_0} \right. \\
&+ \frac{(1 - \beta)^2}{(u + v_0)^2} \sin^2 \frac{an\omega(u + v_0)}{2uv_0} \\
&+ \frac{2(1 - \beta)}{(u^2 - v_0^2)} \sin \frac{an\omega(u - v_0)}{2uv_0} \\
&\cdot \sin \frac{an\omega(u + v_0)}{2uv_0} \cdot \cos \left( \frac{an\omega}{u} + \phi \right) \left. \right\}^{1/2}, \quad (10)
\end{aligned}$$

$$\begin{aligned}
\theta &= \tan^{-1} \frac{\frac{1}{(u - v_0)} \sin \frac{an\omega(u - v_0)}{2uv_0} \cdot \sin \frac{an\omega(u + v_0)}{2uv_0} + \frac{(1 - \beta)}{(u + v_0)} \sin \frac{an\omega(u - v_0)}{2uv_0} \cdot \sin \frac{an\omega(u + v_0)}{2uv_0}}{\frac{1}{(u - v_0)} \sin \frac{an\omega(u - v_0)}{2uv_0} \cdot \cos \frac{an\omega(u + v_0)}{2uv_0} + \frac{(1 - \beta)}{(u + v_0)} \sin \frac{an\omega(u - v_0)}{2uv_0} \cdot \cos \frac{an\omega(u + v_0)}{2uv_0}}. \quad (11)
\end{aligned}$$

The angle of deflection is expressed as  $\alpha$ . It is evident from these equations that the sensitivity is

$$\tan \alpha = \frac{dy}{dx(x=an)} = \frac{dy}{dt} \frac{dt}{dx} = \frac{1}{v_0} \frac{dy}{dt(x=an)} \quad (12)$$

The displacement  $Y$  of the spot on the screen is

$$Y = L \cdot \tan \alpha, \quad (13)$$

where  $L$  = distance from the deflecting device to the screen.

Further considerations on the deflecting angle for special cases will be discussed in the following. When the wave is not reflected at  $Q$ , the deflecting angle  $\alpha$ , can be written as follows by taking the first term of (8); i.e.,

$$\begin{aligned}
\tan \alpha_1 &= \frac{2ekV_m u}{mDv_0\omega} \frac{1}{(u - v_0)} \cdot \sin \frac{an\omega(u - v_0)}{2uv_0} \\
&\cdot \cos \left\{ \omega t - \frac{an\omega(u + v_0)}{2uv_0} \right\}. \quad (14)
\end{aligned}$$

If  $u = v_0$

$$\tan \alpha_2 = \frac{ekV_m an}{mDv_0^2} \cos \left( \omega t - \frac{an\omega}{v_0} \right). \quad (15)$$

It is evident from these equations that the sensitivity is independent of the frequency and is constant over a wide range of frequency.

When  $u \gg v_0$ :

$$\tan \alpha_3 = \frac{2ekV_m}{mD\omega v_0} \sin \frac{an\omega}{2v_0} \cdot \cos \left( \omega t - \frac{an\omega}{2v_0} \right). \quad (16)$$

This equation has the same meaning as the general theoretical formula for the conventional cathode-ray tube. It leads to the same result obtained theoretically by Hollmann.<sup>6</sup>

#### CONSIDERATIONS ON DEFLECTING SENSITIVITY AND EXPERIMENTAL RESULTS

##### 1. Calculated Results of Relationship Between Frequency and Sensitivity when the Electromagnetic Field and Electron Travel in the Same Direction.

Relations between frequency and sensitivity when the electromagnetic field and electron travel in the same direction when reflection does not occur, may be calculated by using (14). In this case, it is assumed that the cosine term which involves the time variable is equal to 1. The phase velocity was chosen to be equal to  $2.93 \times 10^9$  centimeters per second to be in accord with operating conditions.

Fig. 5, curve (a) shows an example when the accelerating voltage is 900 volts (i.e.,  $v_0 = 1.78 \times 10^9$  cm/sec). Sensitivity begins to decrease at a frequency of 400 Mc, and reaches an extremely small value at 2,000 Mc. In the case of  $u > v_0$ , the electron transit time under the deflecting device increases, resulting in the shift of the

<sup>6</sup> H. E. Hollmann, "Die Braunshe Rohre bei sehr Hohen Frequenzen," *Hochfrequenz. und Elektroakust.*, vol. 40, pp. 97-103; September, 1932.

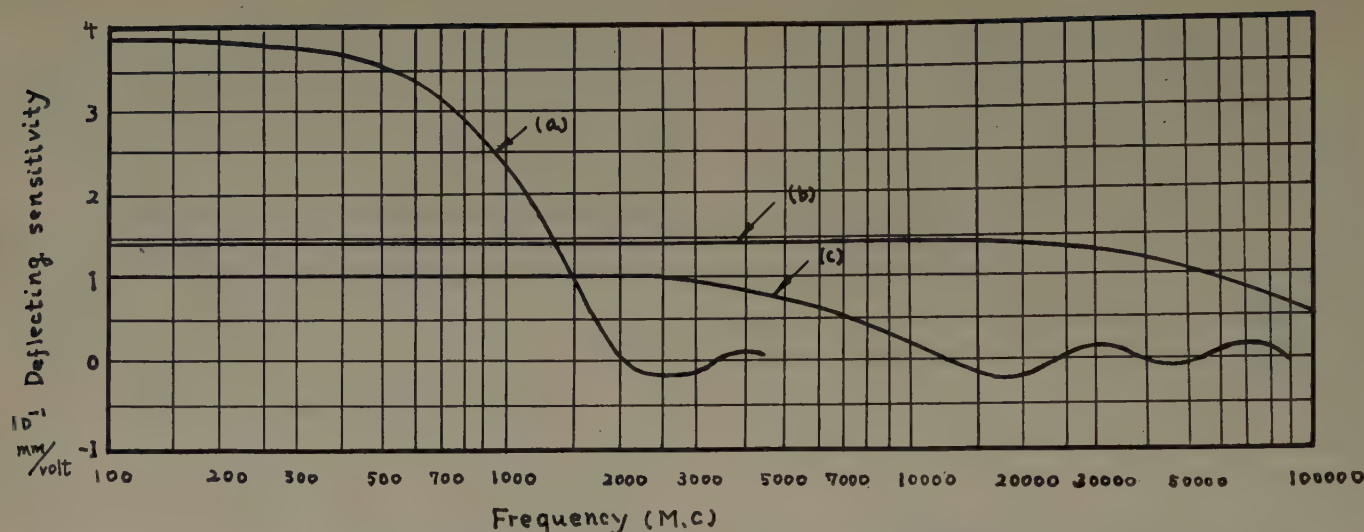


Fig. 5—Relation between frequency and sensitivity when electromagnetic wave and electron travel in same direction. (a) Accelerating voltage, 900 volts; (b) 2,500 volts; (c) 3,025 volts.

falling point of the sensitivity curves to a lower frequency which makes it difficult for application in the uhf range.

Fig. 5, curve (b) shows an example when the accelerating voltage is 2,500 volts (i.e.,  $v_0 = 2.95 \times 10^9$  cm/sec.). Comparing with the former case (a), sensitivity is seen to be lower but the frequency range becomes extremely wide. At 100 kMc ( $\lambda = 3$  mm), sensitivity decreases to 30 per cent. When the phase velocity  $u$  is very nearly equal to  $v_0$ , sensitivity becomes independent of frequency, therefore, uniform measurement from low frequency to uhf is possible in this case. But, it is neglected that the potential distribution along the axis of deflection devices becomes to be departed from cosine distribution when  $l \div \lambda$  ( $l$  of the present tube is 28 mm).

Fig. 5, curve (c) shows an example when the accelerating voltage is 3,025 volts (i.e.,  $v_0 = 3.245 \times 10^9$  cm/sec.). In this case, electron velocity exceeds the phase velocity  $u$ , and the sensitivity decreases.

## 2. Calculated Results of Relationship Between Accelerating Voltage and Sensitivity

Calculations are made on the following two cases: (a) When the electron and electromagnetic wave travel in the same direction without reflection of the wave; and (b) when the wave is reflected at the end of the device and forms a standing wave along the deflecting devices.

Fig. 6 shows the relation between accelerating voltage and sensitivity at frequencies of 500, 1,000, 1,500, and 3,000 Mc in the first case (a). The accelerating voltage at which inversion of deflection occurs will increase with the increase of frequency.

Fig. 7 shows an example at 100 Mc for  $\beta = 0.5$  and  $\phi = \pi/2$ , by using (8) in the second case. In this case, sensitivity decreases to a minimum, different from zero, at 500 volts. When reflection occurs at the end, sensitivity generally does not become zero at any voltage.

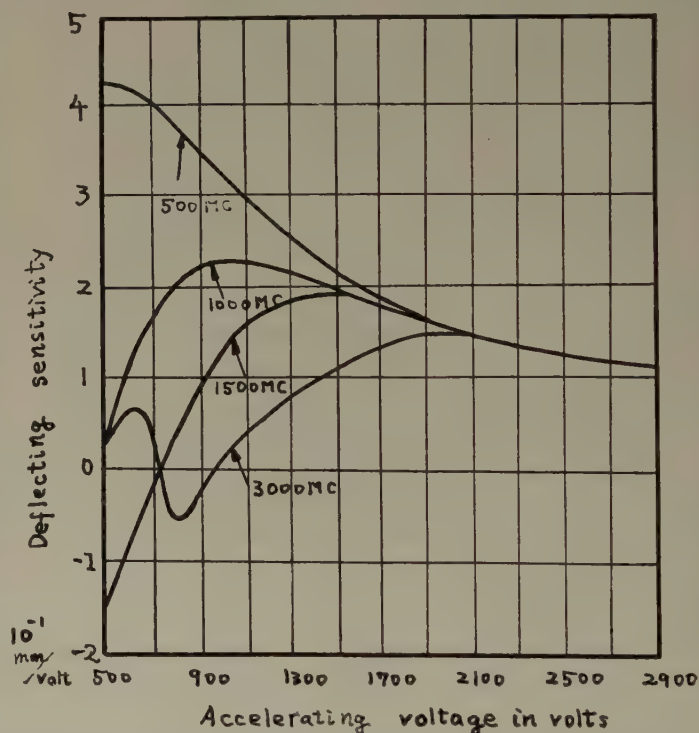


Fig. 6—Relation between accelerating voltage and sensitivity at frequencies of 500, 1,000, 1,500, and 3,000 Mc when electromagnetic wave and electron travel in the same direction.

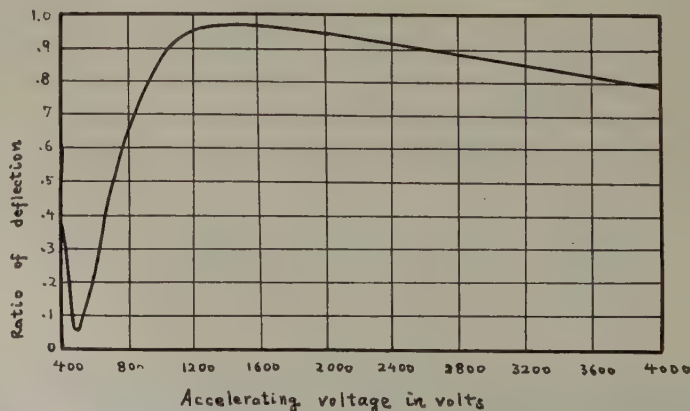


Fig. 7—Relation between accelerating voltage and ratio of sensitivity when the wave is reflected at the end.

### 3. Experiments on the Accelerating Voltage and Sensitivity (By Inversion Spectrum Method)

In order to make clear the relation between the accelerating voltage and sensitivity, Hollmann's method on inversion spectrum has been applied. According to this method, an alternating voltage of 60 cycles is superimposed on the dc accelerating voltage, changing the electron velocity periodically. The alternating voltage of 60 cycles is also supplied to the horizontal deflecting device. An ultra-high-frequency voltage produced by a magnetron oscillator is made to travel along the perpendicular deflecting device. An electron image describing a horizontal line on the screen consists of electron beams of different velocities, in other words, of different accelerating voltages, hence, an inversion spectrum of sensitivity corresponding to each accelerating voltage is obtained on the screen.

Fig. 8 shows the inversion spectrums when the ac rms voltage superimposed on the dc accelerating voltage of 1,040 volts is 370 volts, and a voltage wave of 1,000

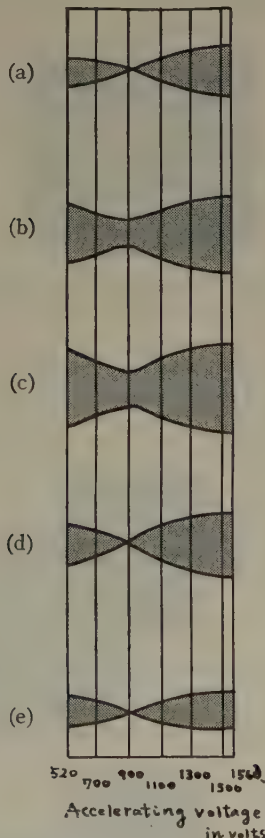


Fig. 8—Inversion spectrum for various terminated conditions. Sections (a) and (e) are cases when the waves are very small. Sections (b), (c), and (d) are cases when reflection occurred.

Mc is propagated along the perpendicular deflecting device. Sections (a), (b), and (c) correspond to the variation of impedance at the end of the perpendicular deflecting device. As is clearly seen, the point of minimum sensitivity and its magnitude varies according to the varia-

tion of the terminating impedance, which is in fair accord with the theoretical deduction previously obtained. The terminal construction of the deflecting device is shown in Fig. 9 in which the terminal impedance may be changed by shifting the tunable stub A.

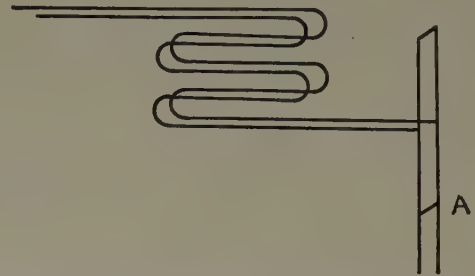


Fig. 9—Illustrated diagram of the deflecting device.

### APPLICATION

#### 1. Observation of Voltage Wave Form

Many difficulties are encountered in the uhf region in obtaining a time axis to observe voltage wave forms. So far, it is impossible to use saw-tooth sweeps in the uhf range, as is done in the low frequencies. As a result, many propositions have been made, among which the the outstanding methods are the following:

- (i) To record the voltage wave by a photograph by using a single sweep time axis.
- (ii) To use the linear portion of a sinusoidal voltage as a time axis.
- (iii) To describe an ultra-dynamical Lissajous figure.

The (i) method requires an extremely rapid sweep in order to make its period equal to the time corresponding to several cycles of the uhf voltage. So rapid is the required velocity of the electron beam across the screen, that its beam density as well as its accelerating voltage must be made large to secure sufficient photographic sensitivity.

But this would result in a decrease of deflection sensitivity. This also requires a long drift space for the electron beam.

The (ii) method implies difficulty in obtaining frequency stability for both the observed voltage and the time axis, whereby, either one of the two should be synchronized with the other to produce a stationary figure on the screen.

The last method (iii) is similar to the usual method of Lissajous figure, wherein, the only difference is that the phase difference between the waves at the horizontal and the vertical deflecting electrodes is controlled by the transit time of the electron beam between the two electrodes.

When the wave involves an  $n$ th harmonic component, the phase difference of the vertical and horizontal voltage affected by the transit time  $\tau$  of the electron is equal to  $\omega \tau$  for the fundamental wave and  $n\omega \tau$  for the  $n$ th harmonic component.

Therefore, the direction of rotation of the fundamental and harmonic vectors has various combinations according to values of  $n$  and  $\tau$ . The most important factor affecting the Lissajous figure is the combination of the direction of the respective vectors incidental to the fundamental and the harmonic component, whether they are unidirectional or antidiagonal. It is sufficient to take into account only these two cases, since the other combinations may be obtained as symmetrical figures of these.

Therefore, the case where the vectors incidental to the fundamental and harmonic wave rotating in the same direction will be considered first. In this example, the direction of rotation of the harmonic vector is coincident with the direction of the movement of the center of the ellipse described by the harmonic vector. This aspect is shown in Fig. 10, where the cycloidal curve has



Fig. 10—Lissajous figure when vectors incidental to fundamental and harmonic wave are rotating in the same direction.

a larger radius of curvature around the outer part than the inner. In general, the inner curve describes a little loop, which will be called a "petal" for convenience. This petal sometimes degenerates into a cusp according to the relative magnitude of the amplitude of the fundamental and the harmonic wave. When the petals are directed toward the interior, as shown in Fig. 10, the figure will be called an "inward figure." Since the harmonic vector rotates  $n$  times while the fundamental describes a single ellipse, the direction of the harmonic ellipse also describes a single rotation resulting in the decrease of the number of the cycloidal loops, which becomes  $(n-1)$ . The photographs shown in Fig. 11 illus-

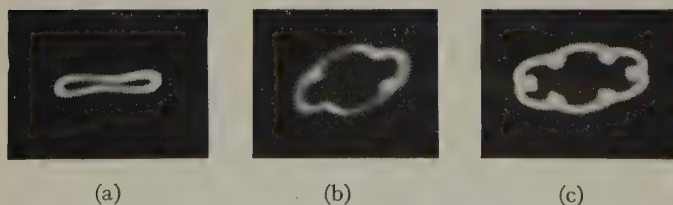


Fig. 11—Photographs of ultradynamic Lissajous figure of distorted waves, involving third (a), fifth (b), and seventh harmonics (c).

trate examples of inward petals of distorted waves involving the third, fifth, and seventh harmonics, respectively.

In cases where the fundamental and harmonic waves rotate in opposite directions, the cycloids are described externally with petals or cusps on the outer sides. This is shown in Fig. 12. The number of petals in this case is



Fig. 12—Lissajous figure when vectors incidental to fundamental and harmonic wave are rotating in opposite direction.



Fig. 13—Photograph of ultradynamic Lissajous figure of distorted wave involving the third harmonic.

equal to  $(n+1)$  Fig. 13 shows examples of such a case when the distorted wave consists of the third harmonic component.

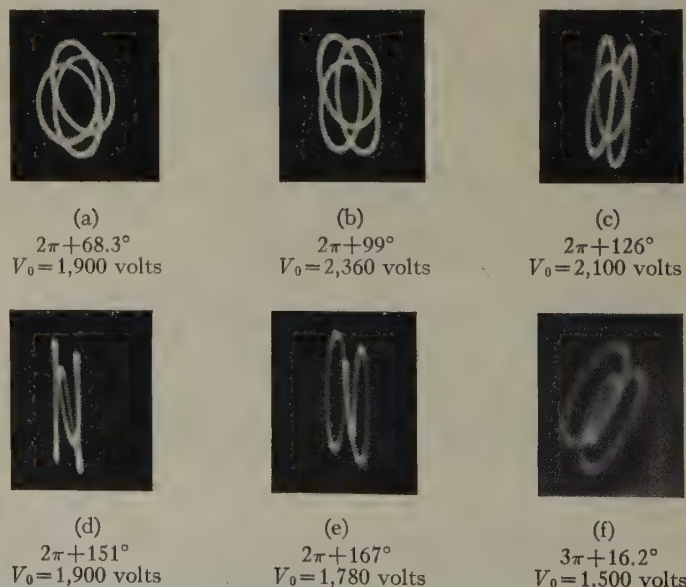


Fig. 14—Photographs of ultradynamic Lissajous figure of distorted wave involving the third harmonic having various phase differences between fundamental and harmonic wave.  $V_0$  = Accelerating voltage of electron beam.

The direction of rotation of the fundamental and the harmonic vector thus varies according to  $\omega\tau$  and  $n\omega\tau$ , which necessarily leads to a frequent variation of the direction of the harmonic vector by the same amount of change in the transit time  $\tau$ . Since it is necessary to consider only the case of unidirectional or antidi-rectional rotation, it is sufficient to take into account the variation rotation of the harmonic wave upon a certain fundamental rotation.

Fig. 14 shows actual examples of such cases, which are the ultradynamic Lissajous figure having the third harmonic at a fundamental wavelength of about 100 cm. Here (a), (b) and (c) are cases of outward petals, (e) and (f) are cases of inward petals. The phase difference of the third harmonic between (a) and (f) is about  $0.7\pi$ .

In the case of the amplitude of the fundamental wave being so small as to be hardly perceptible in the figure, the second, third, or other higher harmonics are apt to be taken as the fundamental, whereby the preceding analysis leads sometimes to harmonics having orders other than numbers. In such a case, the figure is not completed with a single cycle of the apparent funda-

mental, but after several cycles. An example of such procedure is shown in Fig. 15. The order of the harmonics, in such cases, may be determined by the following method.

- (i) Count the number ( $N$ ) of petals.
- (ii) Count the number of petals lying between two successive petals and add one to this number. This is equal to the number of rotation  $P$  of the apparent fundamental wave to return to its original position.
- (iii) The rigorous order of harmonics is equal to:

$$m = \frac{N + P}{P} = n \pm 1,$$

where the negative and the positive signs correspond to the case of outward and inward petals.

An example of such a case is shown in Fig. 16, which is approximately 2.5 times larger than the actual figure. Here,

$$N = 15, \quad P = 4$$

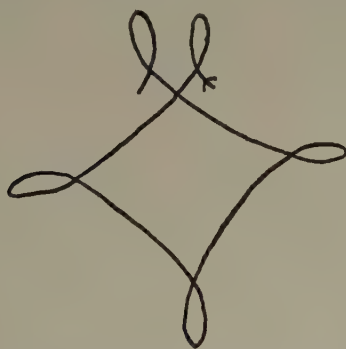


Fig. 15—Lissajous figure involving fractional harmonic.

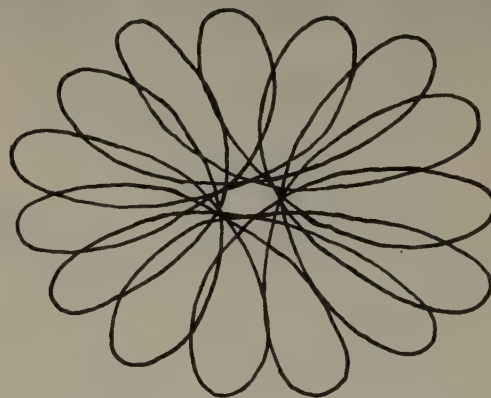
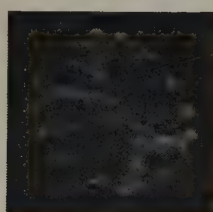


Fig. 16—Ultradynamic Lissajous figure involving fractional harmonic



$$A \sin 4\omega t + B \sin 11\omega t.$$



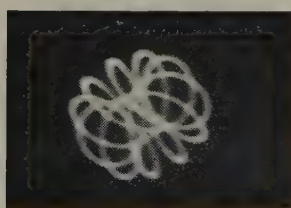
$$A \sin 4\omega t + B \sin 11\omega t.$$



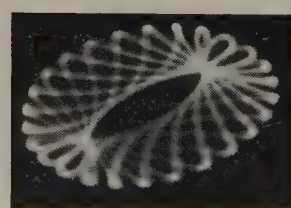
$$A \sin 4\omega t + B \sin 11\omega t.$$



$$A \sin 3\omega t + B \sin 14\omega t.$$



$$A \sin 3\omega t + B \sin 8\omega t.$$



$$A \sin 4\omega t + B \sin 19\omega t.$$

Fig. 17—Photographs of ultradynamic Lissajous figures involving fractional harmonics.

hence,

$$m = \frac{15 - 4}{4} = \frac{11}{4} = 2.75.$$

The original wave may be represented as, (i)  $\sin \omega t + 0.8 \sin 2.75 \omega t$  or (ii)

$$\sum_{n=1}^{\infty} A_n \sin n\omega t, \quad A_n = 0 \text{ except, } A_4 = 1, A_{11} = 0.8.$$

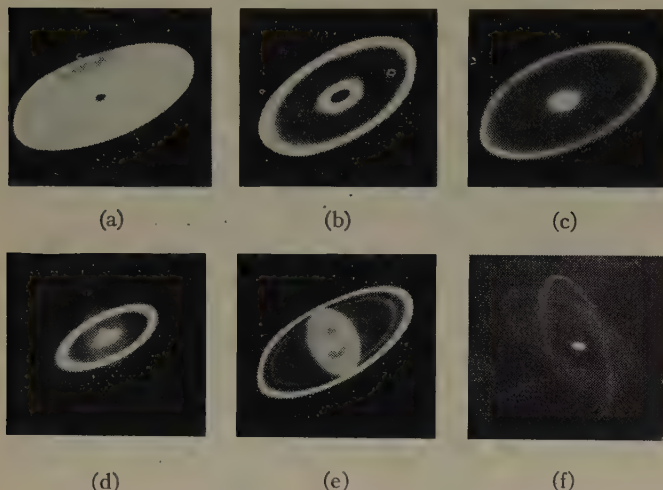


Fig. 18—Photographs of double ellipse obtained with magnetron oscillations.

Fig. 17 shows photographs of ultradynamical Lissajous figure containing the fractional harmonics.

## 2. Measurement of the Degree of Amplitude Modulation

For the measurement of the degree of amplitude modulation using the cathode-ray tube, the trapezoidal or elliptical methods are adopted, in general. The latter is adaptable with the present traveling-wave-type tube.

Fig. 18 shows examples of this method obtained with magnetron oscillations. In Fig. 18(a) is shown the double ellipsoid when the degree of amplitude modulation is about 90 per cent. In this case, the modulation character is nearly linear. In (b) and (c) are shown, respectively, cases of distortion and over modulation. Fig. 18(d), (e) and (f) all show examples of over modulation, but these cases differ from (b), in that oscillations of two different frequencies are represented.

## ACKNOWLEDGMENT

The authors are much indebted to Professors Okabe, Sonoda, and Moto of the Osaka University for their kind advice, and sincere gratitude is due to Messrs. Sakamoto and Otsubo for their valuable assistance in carrying out the experiment. The presentation of this paper is due to H. M. Sarasohn, to whom the authors' special gratitude is expressed.

## Corrections and Additions to

# Speed of Electronic Switching Circuits\*

The authors of the above-mentioned paper have called to the attention of the editors several typographical errors in certain of the equations. The last term of the first equation on page 66 should read as follows:

$$\left[ \frac{1}{R_1 R_2} - g_m^2 \right].$$

The equation (2) gives  $m^2$ , rather than  $m$ .

Equation (4) should read

$$e_2 = \frac{g_m}{2mC_2(m - \alpha)} \left[ \frac{de_1}{dt} \right]_{t=0} e^{(\beta - \alpha)t}.$$

Since this paper deals with an analysis of multivibrators, a footnote reference should be added, calling attention to the work of R. Benjamin,<sup>1</sup> which analyzes blocking oscillators, and in which much the same conclusions are derived.

\* E. M. Williams, D. F. Aldrich, and J. B. Woodford, "Speed of electronic switching circuits," *Proc. I.R.E.*, vol. 38, pp. 65-69; January, 1950.

<sup>1</sup> R. Benjamin, "Blocking oscillators," *Jour. IEE*, vol. 93, Part IIIA, no. 7; 1946.

# Conductivity Measurements at Microwave Frequencies\*

A. C. BECK†, SENIOR MEMBER, IRE, AND R. W. DAWSON†

**Summary**—Because of the skin effect, the surface condition of conductors becomes very important in determining attenuation at microwave frequencies. This has been investigated by measuring small wire samples at a frequency of about 9,000 megacycles. A sample of the wire to be measured is inserted in a metal tube to form the center conductor of an open-ended coaxial line. The ratio of the peak frequency to the half-power bandwidth of this coaxial-line resonator, measured with the aid of an oscillographic display of its amplitude-versus-frequency characteristic, gives its loaded  $Q$ . The amplitude characteristic of the frequency-modulated signal generator, on which a wavemeter marker appears, is viewed simultaneously and used as a reference. By correcting the result to obtain the unloaded  $Q$  of the center conductor alone, the effective conductivity of the sample is obtained.

Results of measurements on a number of samples of different conductors having various surface conditions, treatments, and platings are given. These results are of value in the design of microwave components of all types where loss is a factor of importance.

## I. INTRODUCTION

AT FREQUENCIES in the microwave range, skin effect limits current penetration in conductors to a small fraction of a mil. Therefore the nature and condition of the current-carrying surfaces of microwave equipment have a very important bearing on the attenuation. In order to get quantitative data, equipment was set up to measure the effective conductivity of short pieces of small wire at frequencies in the order of 9,000 megacycles. Various metals, alloys, and plated specimens can be investigated easily and quickly when such small samples are used for the measurements, and it is a simple matter to produce and study the effects of surface roughness, contamination, and corrosion, as well as methods of protection against these conditions.

The equipment described herein was designed to determine the loaded  $Q$  of an open-circuited coaxial line having as a center conductor the sample being measured. This is done by displaying the resonance curve of the coaxial line on an oscilloscope, measuring its peak frequency and half-power bandwidth, and taking their ratio. Effective conductivity is obtained by correcting this result to obtain the unloaded  $Q$  of the center conductor alone.

## II. MEASURING EQUIPMENT

Fig. 1 shows the specimen holder and exciting waveguides used. Small wire specimens, 15 to 20 mils in diameter and less than 5 inches long, are supported in a coaxial line at voltage nulls by polyfoam beads whose dielectric loss is negligible. The ends of the specimen-holder tube are open to permit quick insertion and removal of the samples. This tube is made sufficiently

longer than the sample so that no appreciable radiation losses occur at the open ends, and the ratio of its diameter to that of the wire is large so that its loss is small compared to the loss in the wire. This coaxial line is excited through an orifice from a terminated waveguide by a frequency-modulated signal generator. The output is taken into a second terminated waveguide through an opposite orifice in the coaxial line, and connected to a superheterodyne receiver.

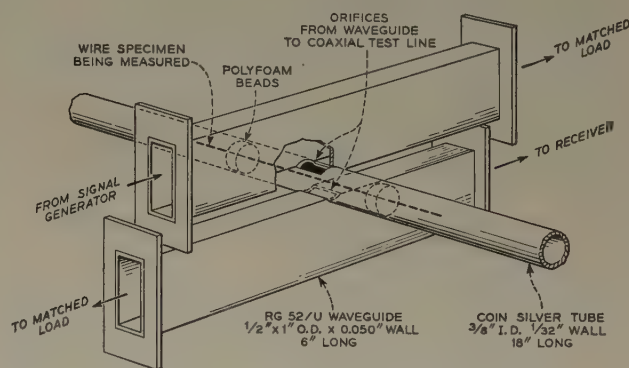


Fig. 1—Specimen holder for wire conductivity measurements.

A block diagram of the measuring equipment is given in Fig. 2, and Figs. 3 and 4 are photographs of the waveguide circuits and the complete equipment. The signal generator, shown at the left center of Fig. 2, and at the extreme left of the photographs, is frequency modulated by placing a fraction of the 60-cycle supply voltage on the repeller in addition to its usual dc voltage. The signal generator output waveguide is connected through a matching and level adjusting attenuator to monitoring directional coupler #1. The forward take-off of this coupler is connected through an adjustable attenuator to a dc-biased crystal detector. The output of this detector is displayed on the oscilloscope, which has a 60-cycle sine-wave sweep of the same phase as the voltage on the oscillator repellers, when the video switching relay is connected on one side. This relay is driven by a 30-cycle oscillator synchronized at the proper phase with the 60-cycle line. A plot of the amplitude-versus-frequency characteristic of the signal generator is shown on the oscilloscope during this time interval. A typical display of this characteristic is shown in Fig. 5(a). The base line is produced when the 30-cycle video switching relay is on the other side, which at this time had no signal.

Following that monitoring branch is directional coupler #2, whose forward take-off is connected through an adjustable attenuator and a cavity wavemeter to another dc-biased crystal detector. The calibration and

\* Decimal classification: R208×R282.1. Original manuscript received by the Institute, January 20, 1950.

† Bell Telephone Laboratories, Inc., Holmdel, N. J.

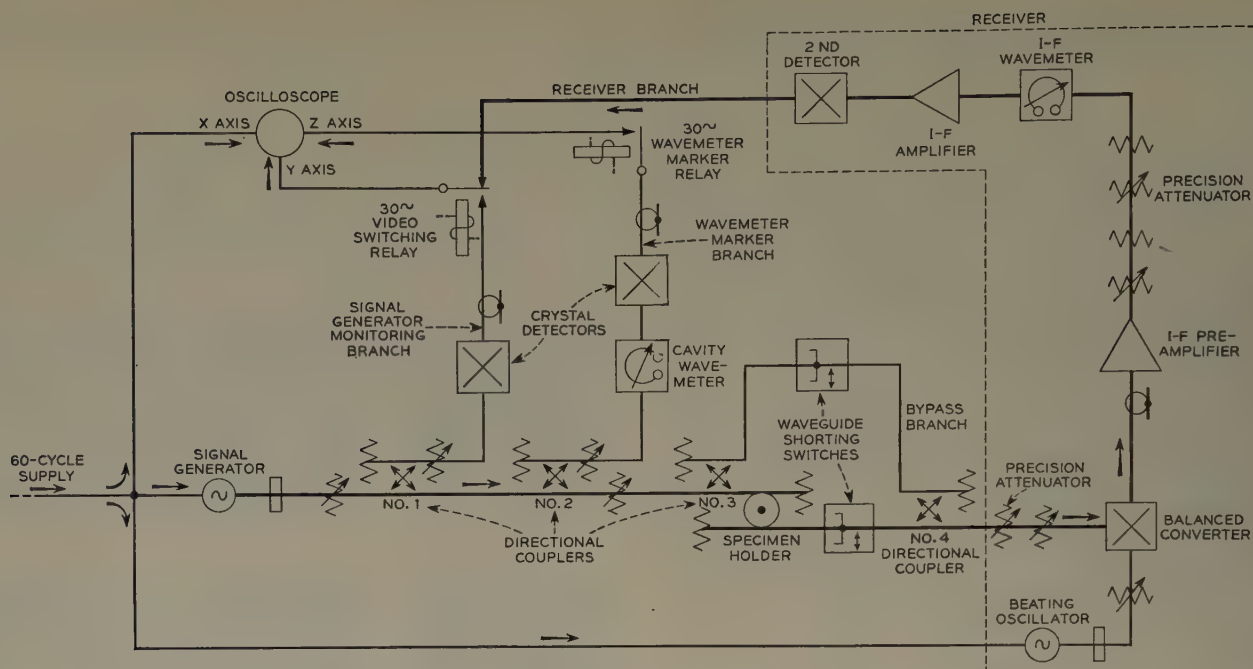


Fig. 2—Block diagram of wire-conductivity measuring equipment.

use of this wavemeter will be discussed later. The output of the detector goes through another 30-cycle relay running in synchronism with the video switching relay to the Z-axis blanking amplifier of the oscilloscope, producing a small gap in the signal-generator monitoring trace corresponding to the frequency at which the cavity wavemeter is set. This gap moves across the face of the oscilloscope as the wavemeter or signal-generator

frequency settings are changed. The appearance of the oscilloscope with the signal generator monitoring detector trace and the frequency marker is shown in Fig. 5(b).

Continuing with Fig. 2, the main signal follows the heavy line through a level adjusting attenuator to directional coupler #3, and then to the specimen holder which has been described and shown in Fig. 1. This is followed by a waveguide shorting switch and directional coupler #4. Between directional couplers #3 and #4 is a separate by-pass waveguide branch paralleling the specimen-holder branch. If the waveguide shorting switch in this by-pass branch is open, and the one in the specimen-holder branch is set to close its waveguide, the

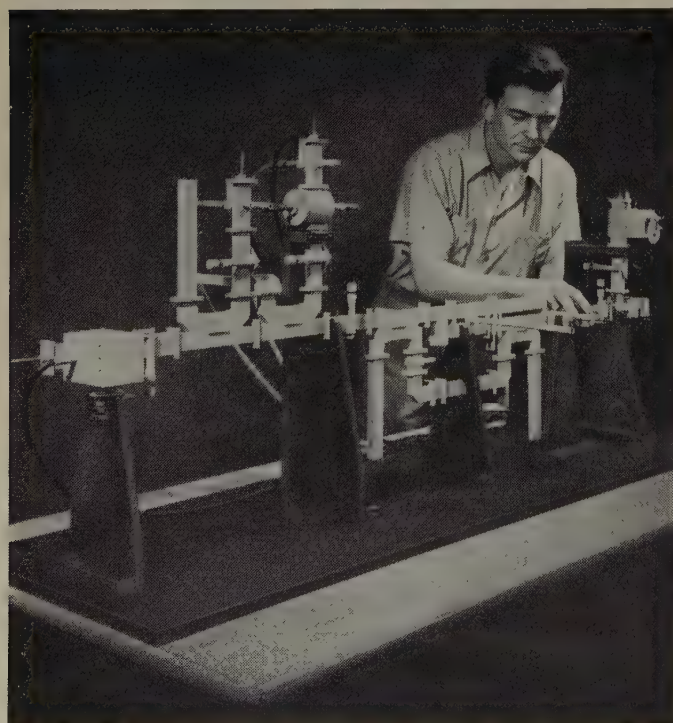


Fig. 3—The waveguide component assembly used for wire-conductivity measurements.

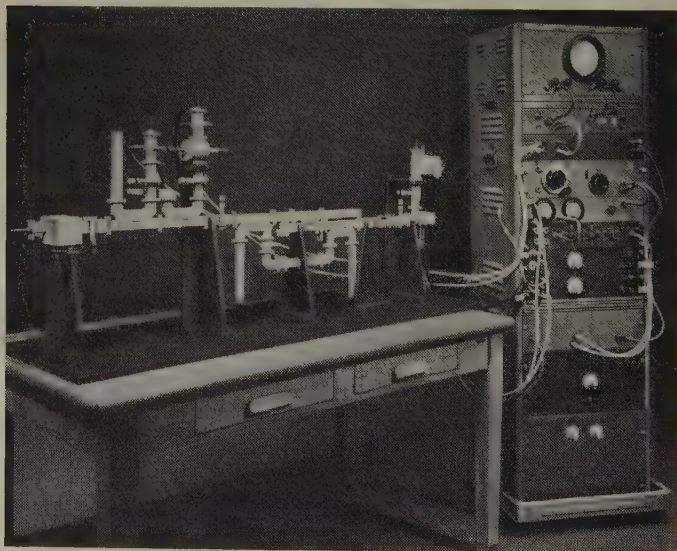


Fig. 4—The complete conductivity-measuring apparatus.

signal goes to the receiver through directional couplers #3 and #4 and the by-pass branch without any frequency-restricting circuits. This switch setting is used for tuning, and as a reference condition for measuring the loss through the specimen holder. When the switch

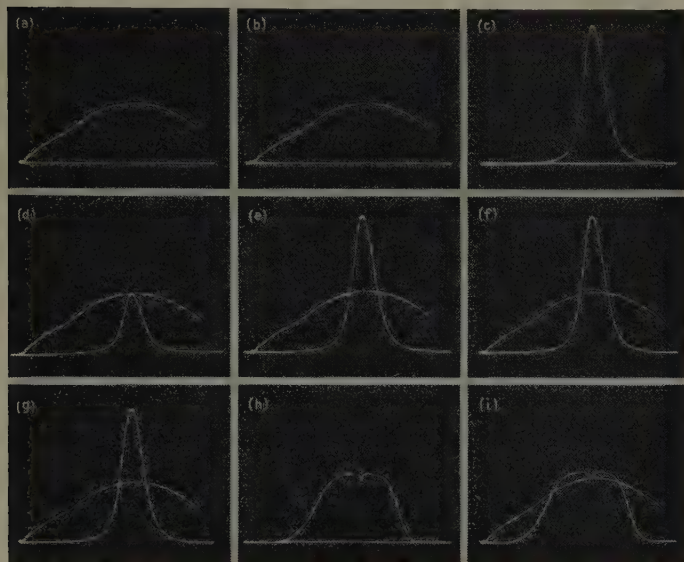


Fig. 5—Photographs of representative oscilloscope traces.

in this by-pass branch is set to block its waveguide path, and the switch in the specimen-holder branch is open, the signal goes to the receiver through the specimen holder, and the resonance characteristic of its coaxial line is displayed so that  $Q$  measurements can be made.

Again referring to Fig. 2, the receiver section is shown in the dashed-line box at the right. After passing through a precision attenuator used for calibration and check purposes, and a level adjusting and terminating attenuator, the signal reaches a balanced crystal converter whose output is the difference between the signal and beating oscillator frequencies. Since the signal is frequency modulated, the beating oscillator is also frequency modulated in synchronism with it, and tuned to a frequency 65 megacycles away from the signal generator frequency. This difference is thus maintained over the frequency-modulation range. The intermediate-frequency output goes to a preamplifier, then to an adjustable calibrated attenuator for measuring the loss in the system, and to the precision attenuator (used for setting the half-power level), which is padded on both sides to maintain a good match. This attenuator is arranged to switch between two positions giving about 8- and 11-db attenuation, respectively, and so adjusted and calibrated that the 3.01-db difference is accurately established. Next is the intermediate-frequency wavemeter, which is an absorption-type instrument with a Cardwell precision capacitor having a 5,000-division scale. It produces a frequency marker pip on the oscilloscope that is used for calibration of the cavity wave-

meter, as will be described later. The main intermediate-frequency amplifier is next, with a gain of about 45 db. This unit, together with the preamplifier, gives an intermediate-frequency amplifier system having a gain of about 85 db, a center frequency of 65 megacycles, a bandwidth of about 20 megacycles one db down, and a low over-all noise figure.

The output of the superheterodyne receiver is displayed on the oscilloscope when the 30-cycle video switching relay is on the opposite side from that used for monitoring the signal generator. A typical trace of this receiver branch signal, with a sample of copper wire in the specimen holder, and no signal on the monitoring branch side of the video switching relay, is shown in Fig. 5(c). A 60-cycle blanking relay is used to produce, from a dc supply, the rectangular wave which is applied to the oscilloscope Z-axis amplifier to blank the return trace, and the ends of the sinusoidal sweep, where the spot is moving slowly. This relay, and the two 30-cycle ones, are Western Electric vibrating-reed mercury-contact relays.

A wave-form timing and phasing diagram is shown in Fig. 6. The 60-cycle oscilloscope blanking relay turns

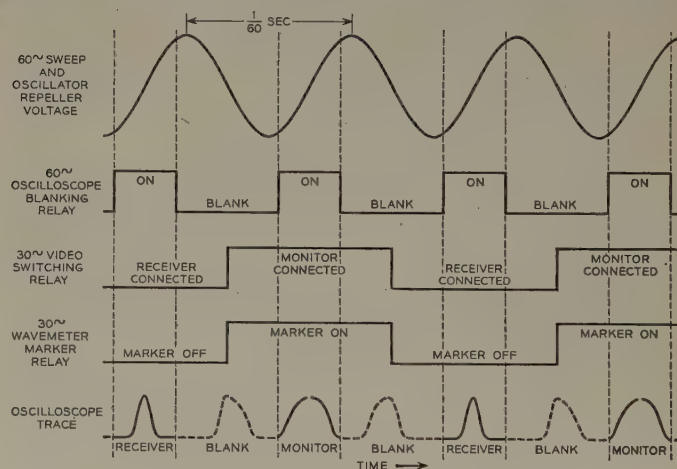


Fig. 6—Wave-form timing and phasing diagram.

on the scope trace during the nearly linear rising voltage part of the 60-cycle line voltage used for sweeping the oscilloscope and oscillators. During this interval the oscilloscope spot moves across the face of the tube from left to right and the frequency of the oscillators goes through its range. The 30-cycle video switching relay alternately connects the scope Y axis to the monitor branch and to the receiver branch, and the cavity wavemeter detector is connected to the oscilloscope Z axis by its relay only when viewing the signal-generator monitor trace. These successive monitor and receiver traces are both seen on the oscilloscope, due to persistence of vision, giving patterns as shown in Fig. 4 ((d), (e), (f), and (g)), which will be explained in more detail later.

### III. AMPLITUDE MEASURING SYSTEM

In order to measure the half-power bandwidth of a

resonant element, means must be provided for accurately determining a 3.01-db difference in signal level. For convenience in use, a special precision attenuator, which was described in Section II, is included in the intermediate-frequency system of the receiver. This attenuator was adjusted to its correct setting with the precision waveguide attenuator, and checked also by direct comparison with intermediate-frequency standard pads that were obtained from Western Electric and checked against standards maintained in these Laboratories. The precision waveguide attenuator, in this case a guillotine-type instrument, has a very low standing-wave ratio. It has a total attenuation of about 4 db, and a long scale dial indicator to measure this range. It can be accurately calibrated on a double detection-measuring receiver against precision intermediate-frequency attenuator pads by measuring the input standing-wave ratio with its output shorted. This method gives a change in intermediate-frequency attenuation several times as large as the change in radio-frequency attenuation, which increases the accuracy of calibration. In addition, in use it is the 3.01-db difference between two settings that is required, and not the absolute value of attenuation at either setting. Differences can be determined more accurately than the absolute values. Using this attenuator to calibrate the intermediate-frequency attenuator also checks the linearity of the equipment between them.

Since the measurement of levels 3.01 db down from the peak of the resonance curve is made at different signal generator frequencies, the change due to its output level variation as a function of frequency must be corrected. This is the reason for displaying on the oscilloscope both the resonance curve of the specimen holder and the amplitude-versus-frequency curve of the signal generator so they are viewed together. Using the latter as the reference line compensates for this change in signal generator output level. There is no parallax difficulty, since these curves are in the same plane. This is, however, only true if the response of the signal-generator monitoring branch and the receiver branch are identical. To make them identical, the monitoring detector and receiver second detector use crystals of the same type that have been selected for matched performance.

At first, even when the receiver was tuned, using the by-pass branch between directional couplers #3 and #4, so that its oscilloscope trace was superimposed on the signal-generator monitor trace,  $Q$  measurements were found to be in error. This error was caused by a change in match and law of response of the receiver second detector with signals having different dc components, which is the case when switching from the by-pass branch for tuning to the specimen-holder branch for making measurements. This difficulty was overcome by using a large dc bias on the second detector to maintain a constant input impedance and to swamp out the

small dc component changes with the different signals. To keep the detector characteristics alike, and to obtain a good match, dc bias is also used on the monitoring detectors. To maintain alignment of the base lines when these different signals are displayed, direct-coupled video amplification is used in the vertical deflection system of the oscilloscope. Initial alignment of the base lines is obtained by small adjustments in the detector bias voltages.

#### IV. FREQUENCY MEASURING SYSTEM

The  $TE_{011}$  cavity-type double-orifice coupled wavemeter mentioned before produces the frequency marker used for measuring the absolute frequency and the half-power bandwidth. It has a high enough  $Q$  to give a sharp marker, as can be seen in Fig. 5. Frequency readings are obtained from the calibration of the micrometer which drives the piston, and for measuring bandwidth, a fine control was added at the opposite end of the instrument, as can be seen from the photographs, Figs. 3 and 4. This consists of a movable silver rod of small diameter extending into the cavity. A dial spread of about forty divisions per megacycle is obtained on the micrometer driving this rod.

Since the calibration of this fine-control micrometer is not linear, and also is a function of the setting of the main piston, it is calibrated each time it is used. The first step in this calibration is to remove the frequency modulation from the receiver beating oscillator by disconnecting the 60-cycle voltage on its repeller. Since the signal generator is still frequency modulated, the frequency response curve of the receiver, modified by the signal-generator amplitude characteristic, is now traced out on the oscilloscope. The precision intermediate-frequency absorption wavemeter is used to put a downward pip on this trace, as shown on Fig. 5(h). This intermediate-frequency wavemeter has about 200 divisions per megacycle, and has been accurately calibrated with a heterodyne frequency meter which is checked with station WWV. Settings of the high-frequency cavity wavemeter are transferred to this wavemeter by aligning the intermediate-frequency wavemeter pip on the receiver trace with the frequency marker on the monitor trace, as shown in Fig. 5(i). After doing this for the two settings of the cavity wavemeter obtained at the half-power points on the resonance curve, the bandwidth is obtained by subtracting the two frequency readings of the intermediate-frequency wavemeter.

This method of transferring the readings to the intermediate-frequency wavemeter can be in error if there is any residual or pickup frequency modulation on the beating oscillator of the receiver. To avoid this possibility, dc is used on the heater of the oscillator tube, all leads are well shielded, and the repeller is grounded for ac by a large capacitor when frequency modulation is removed for this calibration. To make sure that this was effective, an electron-coupled video-frequency oscil-

lator known to have no frequency modulation was connected to drive the crystal in the detector of the cavity wavemeter branch. This modulated the high-frequency signal, so the cavity wavemeter appears to give three markers instead of one. The middle one occurs when the signal generator sweeps across the resonant frequency of the cavity wavemeter, and the others are the sidebands due to its amplitude modulation, so the spacing between the two outside markers is twice the checking video oscillator frequency. To check for beating-oscillator frequency stability, the spacing of these markers was measured with the intermediate-frequency wavemeter. The results showed that beating-oscillator frequency modulation was negligible when calibrating the cavity wavemeter with this equipment.

## V. MEASUREMENT PROCEDURE

The waveguide switches are set for the signal to pass through the specimen holder. After inserting a sample several half wavelengths long, the signal generator and receiver are tuned approximately to locate a resonance. The sample wire is adjusted in lengthwise position to give maximum signal transmission. Then the waveguide shorting switches are reversed so that the signal goes to the receiver through the by-pass branch, and the beating-oscillator frequency and sweep magnitude are carefully adjusted to superimpose the receiver and monitor traces on the oscilloscope. The receiver is then connected through the specimen holder again by means of the waveguide shorting switches, and adjusted to a level which gives trace alignment as shown in Fig. 5(d). The power transmission ratio through the specimen holder at resonance is measured with the intermediate-frequency attenuator by comparing these last two conditions, and correcting for the fixed loss through the directional coupler path, which has been previously determined. The frequency of the resonance peak is measured with the cavity wavemeter marker using the main dial, with the fine frequency micrometer set at its fixed reference position. The attenuator next to the signal generator is varied to check that the peak of the resonance-curve trace stays superimposed on the monitor trace as the level changes. This is important, as it is a check on the linearity of the system. Then the precision intermediate-frequency attenuator is changed to its 3.01-db lower setting, which changes the pattern to the one shown in Fig. 5(e). Using an optical magnifier on the oscilloscope, the half-power bandwidth is measured with the cavity-wavemeter fine frequency dial by putting the marker on each side as shown in (f) and (g) Fig. 4. Calibration of these readings with the intermediate-frequency wavemeter as previously described gives the frequency bandwidth. One half wavelength is then cut off the sample and the new length is measured. This process is repeated for several multiples of a half wavelength.

## VI. CORRECTIONS AND CALCULATIONS

The measured loaded  $Q$  for each multiple half wavelength of line is obtained by taking the ratio of the resonant frequency to the half-power bandwidth. Different sample lengths are usually measured at slightly different resonant frequencies, because in practice it is difficult to cut exact half wavelengths off the specimen. For this reason, the  $Q$  values for each length are converted to values at a convenient common frequency by use of the relation that  $Q$  is inversely proportional to the square root of wavelength. These values still include the effects of losses through the orifices of the coaxial test line into the waveguide system, the end effects due to fringing fields, and the outer conductor losses.

The unloaded  $Q$  values, which would have been measured if there had been no losses through the coupling orifices, can be obtained from the loaded  $Q$  values by using the measured power-transmission ratio through the specimen holder at resonance. Denoting the unloaded  $Q$  by  $Q_0$ , the loaded  $Q$  by  $Q_l$ , and the power-transmission ratio at resonance by  $T$ , this correction is made by using the relation

$$Q_0 = \frac{Q_l}{1 - 2\sqrt{T}} \quad (1)$$

which is derived in the appendix. Although each wire is inserted as near as possible to the position giving greatest transmitted power at resonance, exact location is difficult, but since the measurements are made for a fixed wire position, this correction compensates for small misalignments of the specimen with respect to the orifices. With the specimen holder used in this work, the measured power-transmission ratio at resonance is between 30 and 40 db, so this amounts to a correction of two to six per cent of the  $Q$  values. Smaller orifices would mean a smaller correction, but also less power transmitted, so noise on the receiver trace would be more of a problem. The present design seems to represent a good compromise in this respect. With no center conductor wire in the specimen holder, the output is about 90 db below the input, so no correction is necessary for leakage power.

Measurements are made on several multiple half wavelengths of wire in order to increase the accuracy by averaging a number of readings, to correct for end effects, and to find nonhomogeneous samples. From the wire lengths and the frequencies, the number  $n$  of half wavelengths of each measured piece of the sample is determined. A plot of  $n\lambda/2Q_0$  for each length versus  $n\lambda/2$  is then made. An example of such a plot is shown in Fig. 7. This should be a straight-line relation, so departures from linearity check sample homogeneity and random errors of measurement. End effects in the coaxial resonator are eliminated by taking the slope of this plot. From the slope of a straight line drawn as carefully as possible through the plotted points, the

total unloaded  $Q$  of the coaxial line is determined, as shown in Fig. 7, and denoted by  $Q_t$ .

The effect of losses in the outer conductor must be taken out, since we are interested in the losses due to the center conductor alone. The formulas for the  $Q$  of a coaxial line, which can be obtained from numerous sources,<sup>1</sup> show how these can be separated. It is possible to write

$$\frac{1}{Q_t} = \frac{1}{Q_a} + \frac{1}{Q_b} \quad (2)$$

where  $Q_t$  is the total unloaded  $Q$  of the coaxial line, and  $Q_a$  and  $Q_b$  are defined as  $Q$  values corresponding to the inner conductor and the outer conductor, respectively. For the specimen holder used in this work the outer conductor loss is only two to five per cent of the total loss, so the required accuracy of determining outer conductor loss is not very high. It was determined by assuming its value in the first place, then measuring a specimen wire of the same material with as near as possible the same surface conditions, and using the first assumption in correcting its measurement. Using the more accurate figure obtained by this measurement for the material, instead of the assumed one, the outer conductor correction was again made, and a new result obtained. Such a procedure converges rapidly to a result of more than the required accuracy. For correcting all the measurements obtained with this equipment,  $Q_b$  is obtained from this result and the relation that it is inversely proportional to the square root of wavelength.  $Q_a$  is then obtained from  $Q_t$  by means of (2).

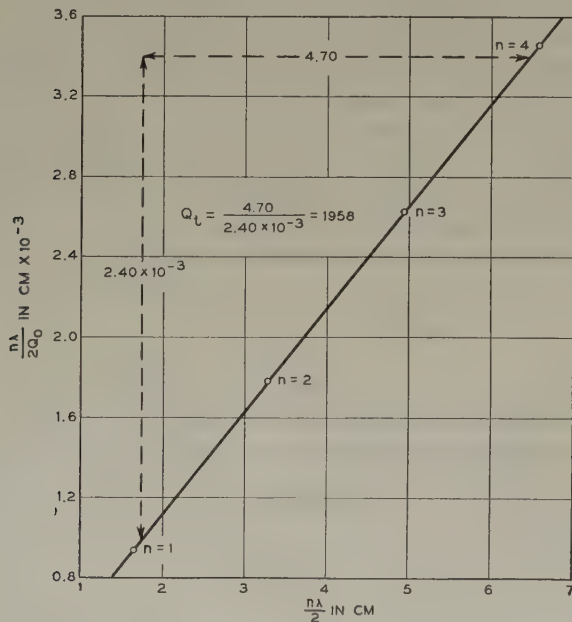


Fig. 7.—A sample plot of measurements on a specimen of vacuum annealed copper wire, showing how  $Q_t$  is obtained from the slope of the line.

<sup>1</sup> For example, S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., Inc., New York, N. Y., Sections 4.9, 7.22 and 8.8; 1943.

The results of measurements with this equipment are usually given in terms of  $Q_a$ , or its ratio to theoretical values, or to values for other materials or conditions. Such quantities are given in Table I. A measurement of  $Q_a$  with this equipment by this method takes about two hours to make.  $Q_a$  is inversely proportional to the attenuation of microwave components using the same material in the same condition. However, formulas for computing the attenuation of most components, particularly coaxial lines or waveguides, are usually given in terms of the conductivity of the material of which they are made. The formula relating  $Q_a$  to the effective conductivity is given by Schelkunoff<sup>1</sup> in terms of the "intrinsic resistance"  $R$  as follows:

$$Q_a = \frac{4\pi^2 K}{R\lambda/a} \quad (3)$$

where

- $K$  = surge impedance of line =  $138.155 \log_{10} b/a$
- $a$  = outer radius of center conductor in centimeters
- $b$  = inner radius of outer conductor in centimeters
- $\lambda$  = free-space wavelength in centimeters.

This intrinsic resistance is related to the conductivity by the formula

$$R = \frac{344.144\sqrt{\mu}}{\sqrt{g}\sqrt{\lambda}} \quad (4)$$

where

- $\mu$  = permeability relative to free space
- $\lambda$  = free-space wavelength in centimeters
- $g$  = conductivity in mhos per meter.

The intrinsic resistance and the effective conductivity of any sample can be obtained from  $Q_a$  by means of (3) and (4).

In order to relate the measurements to calculated values, the theoretical  $Q_a$  can be calculated from the measured dc resistance of the sample wires. In making this measurement of dc resistance, a long piece of the sample wire was used when possible in the Wheatstone Bridge, and its length and temperature were also recorded. Its dc resistance at 20° C was obtained by means of the usual relation

$$r_{20} = \frac{r_t}{1 + \alpha(t - 20)} \quad (5)$$

where

- $r_t$  = total measured dc resistance of the wire
- $\alpha$  = temperature coefficient of resistance
- $t$  = temperature of wire in degrees C.

From this, the relative conductivity  $C$ , with respect to the International Annealed Copper Standard, is calculated from the formula

$$C = \frac{0.15328L}{88900Ar_{20}} \quad (6)$$

TABLE I

Material	C	Diameter (Mils)	$\lambda$ cm	$Q_s'$	$Q_s''$	$Q_s$	$Q_s/Q_s''$	$Q_s/Q_s'$	Remarks
Silver	1.0629	19.8	3.272	2202	2145	1927	0.90	0.87	Commercial wires
Copper	1.0293	19.6	3.280	2144	2113	1880	0.89	0.88	
24K Gold	0.7634	20.0	3.283	1881	2155	1794	0.83	0.95	
Aluminum	0.6106	19.4	3.277	1651	2113	1469	0.69	0.89	
Tungsten	0.3128	19.8	3.259	1201	2148	1011	0.47	0.84	
Molybdenum	0.3095	20.5	3.284	1219	2188	1160	0.53	0.95	
Platinum	0.1581	20.0	3.286	896	2155	898	0.42	1.00	
Coin Silver	0.8418	20.0	3.280	1977	2155	1705	0.79	0.86	Commercial alloys
Soft Brass	0.2700	19.8	3.287	1116	2139	1069	0.50	0.96	
Phosphor Bronze	0.1498	20.0	3.267	832	2160	811	0.38	0.97	
Silver	1.0832	20.0	3.264	2248	2160	1894	0.88	0.84	Vacuum-annealed metals and alloys
Copper	1.0490	19.5	3.296	2164	2103	2038	0.97	0.94	
24K Gold	0.7560	19.5	3.281	1842	2108	1707	0.81	0.93	
Aluminum	0.6214	19.6	3.265	1680	2118	1611	0.76	0.96	
Molybdenum	0.3095	19.5	3.270	1180	2112	1133	0.54	0.96	
Spring Brass	0.2246	19.6	3.300	997	2097	1006	0.48	1.01	
Phosphor Bronze	0.1347	19.6	3.264	828	2118	809	0.38	0.98	
Iron	0.1873	15.3	3.264	788	1802	297	0.16	0.38	Commercial annealed
Silver	1.0629	19.0	3.272	2145	2090	1987	0.95	0.93	Commercial wires hand-polished with #3 alumina, rubbed lengthwise only; 100 strokes
Copper	1.0293	19.2	3.284	2127	2096	2009	0.96	0.95	
24K Gold	0.7634	20.0	3.268	1883	2160	1875	0.87	1.00	
Aluminum	0.6106	19.0	3.273	1630	2090	1492	0.71	0.92	
Molybdenum	0.3095	20.5	3.274	1220	2191	1198	0.55	0.98	
Phosphor Bronze	0.1498	20.0	3.270	832	2159	840	0.39	1.01	
Soft Brass	0.2700	19.5	3.286	1099	2107	1086	0.52	0.99	
Copper <sup>1</sup>	1.0293	19.9	3.282	2178	2147	844	0.39	0.39	Threaded 220 threads per inch
Soft Brass <sup>1</sup>	0.2700	18.8	3.290	1073	2063	1025	0.50	0.96	Electropolished
Copper <sup>1</sup>	1.0293	19.2	3.277	2129	2098	1928	0.92	0.91	Electropolished
Copper <sup>1</sup>	1.0293	19.2	3.280	2127	2096	2068	0.99	0.97	Smooth electropolished
Copper <sup>1</sup>	1.0293	19.2	3.284	2127	2096	2088	1.00	1.00	Very smooth electropolished
Copper <sup>2</sup>	1.0293	19.2	3.277	2129	2098	1928	0.92	0.91	Immediately after electropolishing
Copper <sup>2</sup>	1.0293	19.3	3.266	2139	2108	1940	0.92	0.90	2 hours after lacquering
Copper <sup>2</sup>	1.0293	19.8	3.288	2170	2139	1982	0.93	0.91	12 days after electropolishing
Copper <sup>2</sup>	1.0293	19.4	3.271	2145	2114	1983	0.94	0.92	12 days after lacquering
Copper <sup>2</sup>	1.0293	19.0	3.265	2121	2091	1843	0.88	0.87	33 days after electropolishing
Copper <sup>2</sup>	1.0293	19.1	3.2810	2120	2090	1897	0.91	0.89	33 days after lacquering
Copper <sup>2</sup>	1.0293	19.4	3.2638	2148	2117	1879	0.89	0.87	64 days after electropolishing
Copper <sup>2</sup>	1.0293	19.8	3.2783	2173	2142	1973	0.92	0.91	64 days after lacquering
Copper <sup>2</sup>	1.0293	19.2	3.2737	2130	2097	1990	0.94	0.93	3 months after lacquering
Copper <sup>2</sup>	1.0293	20.0	3.2697	2190	2158	1932	0.89	0.88	5 months after electropolishing
Copper <sup>3</sup>	1.0293	18.8	3.273	2100	2070	1955	0.94	0.94	20 lengthwise strokes #4/0 emery; grooves 1 micron wide, 1 micron deep
Copper <sup>3</sup>	1.0293	18.7	3.298	2083	2052	1830	0.89	0.88	20 lengthwise strokes #2/0 emery; grooves 2.6 microns wide, 2 microns deep
Copper <sup>3</sup>	1.0293	18.5	3.293	2074	2044	1578	0.77	0.76	20 lengthwise strokes #1 emery; grooves 5.4 microns wide, 4.5 microns deep
Copper <sup>3</sup>	1.0293	18.6	3.289	2080	2050	1218	0.59	0.59	Rolled between ground steel plates producing circumferential grooves, 4 microns wide, 3.5 microns deep
Copper <sup>3</sup>	1.0293	19.2	3.284	2127	2096	2088	1.00	0.98	Immediately after electropolishing
Copper <sup>3</sup>	1.0293	19.0	3.284	2115	2085	2033	0.97	0.96	3 days after electropolishing
Copper <sup>3</sup>	1.0293	19.2	3.297	2123	2093	2012	0.96	0.95	10 days after electropolishing
Copper <sup>3</sup>	1.0293	18.9	3.274	2104	2074	2010	0.97	0.95	13 days after electropolishing
Copper <sup>3</sup>	1.0293	18.5	3.257	2086	2060	1938	0.94	0.93	21 days after electropolishing
Copper <sup>3</sup>	1.0293	19.0	3.270	2116	2088	2029	0.97	0.95	41 days after electropolishing
Copper <sup>3</sup>	1.0293	19.0	3.262	2115	2087	1891	0.91	0.89	18 weeks after electropolishing
Copper <sup>3</sup>	1.0293	20.0	3.281	2185	2155	1561	0.72	0.71	Aged out of doors, 4½ months
Copper <sup>4</sup>	1.0	20.8	3.298	2204	2204	1814	0.82	0.82	Acid bath 20 minutes at 200 ma
Copper <sup>4</sup>	1.0	20.1	3.294	2158	2158	1787	0.83	0.83	Acid bath 20 minutes at 200 ma, then hand-polished 100 strokes #3 alumina, lengthwise, only
Copper <sup>4</sup>	1.0	20.0	3.291	2152	2152	1683	0.78	0.78	Acid bath 30 minutes at 100 ma
Copper <sup>4</sup>	1.0	20.8	3.265	2215	2215	2100	0.95	0.95	Acid bath 20 minutes at 200 ma, then re-electropolished with same process that produced 1.0 $Q_s/Q_s''$
Copper <sup>4</sup>	1.0	20.8	3.263	2215	2215	1874	0.85	0.85	5 months exposure to air; same treatment as previous sample

TABLE I—(continued)

Material	C	Diameter (Mils)	$\lambda$ cm	$Q_a'$	$Q_a''$	$Q_a$	$Q_a/Q_a''$	$Q_a/Q_a'$	Remarks
Silver <sup>5</sup>	1.0843*	20.9	3.275	2311	2229	1019	0.46	0.44	Cyanide 15 minutes at 80 ma
Silver <sup>5</sup>	1.0843*	19.8	3.272	2238	2143	1268	0.59	0.57	Cyanide 6 minutes at 100 ma
Silver <sup>5</sup>	1.0843*	19.8	3.278	2236	2141	1628	0.76	0.73	Cyanide 25 minutes at 25 ma
Silver <sup>5</sup>	1.0843*	19.8	3.265	2239	2143	1498	0.70	0.67	Cyanide 25 minutes at 25 ma, hand-polished #3 alumina; 100 strokes lengthwise

\* Assumed relative conductivity of silver.

<sup>1</sup> Commercial wires.<sup>2</sup> Electropolished copper wire of 0.92  $Q_a/Q_a''$ .<sup>3</sup> Electropolished copper wire of 1.0  $Q_a/Q_a''$ .<sup>4</sup> Copper-electroplated on electropolished copper of 1.0  $Q_a/Q_a''$ .<sup>5</sup> Silver-electroplated on electropolished copper of 1.0  $Q_a/Q_a''$ .

where

$L$  = wire length in centimeters

$A$  = wire area in square centimeters.

The conductivity  $g$ , in mhos per meter, is given by the relation

$$g = 5.8005 \times 10^7 \times C. \quad (7)$$

This can be used in (4) to get the theoretical intrinsic resistance, and, from this, (3) gives the theoretical value of  $Q_a$  calculated from the measured dc resistance.

## VII. RESULTS

Table I gives results of measurements with this equipment for a number of specimen wires. The first column lists the material under test. The second column, headed "C," gives the relative conductivity of the sample as calculated by (5) and (6) from dc resistance measurements. Next are listed the diameter of the sample and the wavelength of the measurement. The column headed  $Q_a'$  gives the theoretical  $Q_a$  value calculated from this dc conductivity, diameter, and wavelength, as explained in the previous section. The column headed  $Q_a''$  gives the theoretical  $Q_a$  calculated for copper of unit relative conductivity, and the given diameter and wavelength. The column headed  $Q_a$  gives the value measured at the given wavelength with this equipment. The next column gives the ratio  $Q_a/Q_a''$ , a quantity which is useful in comparing the attenuation of microwave systems made of this material with the theoretical attenuation using standard copper. The increase in attenuation is obtained by dividing by this number, since attenuation is inversely proportional to  $Q$ . The last column gives the ratio  $Q_a/Q_a'$ , which indicates the relation between the measured results and calculated values based on the dc-resistance measurements of the same samples.

As a check on the accuracy of measurement with this equipment, it will be seen that very smooth samples of platinum, phosphor bronze, brass, molybdenum, 24K gold, and copper gave results within 2 per cent of the  $Q$

calculated from measurements of their dc resistance. These metals have  $Q_a$  values in the range from about 800 to 2,100 when used in this specimen holder.

The data of Table I show quantitatively the importance of a smooth surface, especially for high conductivity samples, at these frequencies where the current skin depth is in the order of one thirtieth of a mil for copper. Microscopic examination showed that the hand-polished samples were still not very smooth. Lengthwise grooves or scratches do not reduce the effective conductivity as much as circumferential ones, as would be expected. Theoretically a smooth 60° V thread on the surface should reduce the  $Q$  to one half its value for a smooth sample,<sup>2</sup> but the measurement of a rough thread shows a reduction to about 40 per cent of the smooth sample value.

It will be noted that electropolishing is somewhat variable, but that good results can be obtained. Microscopic examination of the surface always showed good correlation between smoothness and measured high conductivity. The effective conductivity of freshly electropolished copper is seen to drop in value due to aging in the laboratory, and to drop further if aged outdoors, as would be expected. To protect a freshly polished copper surface, DuPont clear lacquer was selected, because it was considered to be good for this purpose, and was found to cause a negligible increase in loss at these frequencies.

Fragmentary measurements on electroplated samples gave rather poor results. Since plated metal is usually more rough and porous than solid metal, this would be expected. It will be noted that copper-plated samples could not be electropolished to as high a conductivity as solid samples, probably because of the lower density of the plating due to its porosity. Silver plated on very smooth electropolished copper samples from a cyanide bath gave very poor results, with the lowest values for rough porous platings at high current densities.

<sup>2</sup> S. P. Morgan, Jr., "Effect of surface roughness on eddy current losses at microwave frequencies," *Jour. Appl. Phys.*, vol. 20, pp. 352-362; April, 1949.

## VIII. ACKNOWLEDGMENTS AND CONCLUSIONS

We have received suggestions and encouragement from many of our colleagues, particularly H. T. Friis and A. E. Bowen, under whose supervision this work was done, and W. A. Tyrrell, who in addition helped with the plating, chemical treatment, and microscopic examination of specimens. Assistance in the measurements was given by G. D. Mandeville.

The information obtained with this measuring equipment has been found useful in the design and treatment of microwave components of all types where loss is a factor of importance. The values obtained in this study of small wires agree very well with the results of waveguide attenuation measurements made by other methods for several materials. The versatility, rapidity, and accuracy of measurements by this method have made this a valuable tool in microwave work.

## APPENDIX

*Derivation of the Relation between the Unloaded  $Q$  and the Loaded  $Q$  of the Coaxial Line in Terms of the Transmission of Energy through the Specimen Holder*

Fig. 8 gives a representation of the equivalent circuit of the specimen holder. The orifices between the coaxial line and the waveguides are depicted by ideal trans-

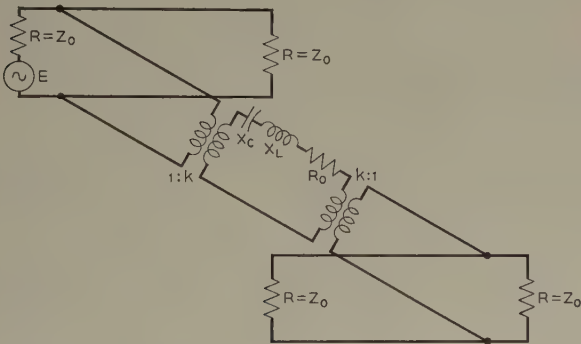


Fig. 8—A representation of the equivalent circuit of the specimen holder shown in Fig. 1.

formers, here assumed to be identical. Fig. 9 shows this circuit transformed into the upper mesh. The measuring method determines the loaded  $Q$ , here called  $Q_l$ , of this circuit. This is the ratio of inductive reactance to total equivalent series resistance in the circuit. Hence

$$Q_l = \frac{X_L/k^2}{R/2 + R_0/k^2 + R/2} = \frac{X_L}{k^2R + R_0}. \quad (8)$$

We need to know the unloaded  $Q$  of the coaxial line itself, the value it would have if there were no losses

through the orifices into the rest of the circuit. We denote this by  $Q_0$ . This is the ratio of inductive reactance to the equivalent series resistance of the coaxial line itself. Hence

$$Q_0 = \frac{X_L/k^2}{R_0/k^2} = \frac{X_L}{R_0}. \quad (9)$$

Taking the ratio of (9) to (8)

$$\frac{Q_0}{Q_l} = \frac{k^2R + R_0}{R_0}. \quad (10)$$

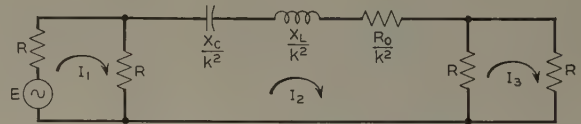


Fig. 9—The circuit of Fig. 8 transformed into the upper mesh.

In order to determine this ratio, the power transmission through the specimen holder at resonance is measured. This is the ratio of the power output from the specimen holder to the power available at its input, in both cases between matched loads at the generator and receiver. Accordingly, as defined above, denoting the power transmission by  $T$ , the output power by  $P_0$ , and the available power by  $P_A$

$$T = P_0/P_A. \quad (11)$$

From Fig. 9

$$P_0 = I_3^2 R \quad (12)$$

and

$$P_A = E^2/4R. \quad (13)$$

Solving the network of Fig. 9 at resonance for  $I_3$  in terms of  $E$  gives

$$I_3 = \frac{k^2 E}{4(k^2 R + R_0)}. \quad (14)$$

Putting (14) into (12), and that result and (13) into (11) gives

$$T = \frac{k^4 R^2}{4(k^2 R + R_0)^2}. \quad (15)$$

Combining (10) and (15) gives

$$\frac{Q_0}{Q_l} = \frac{1}{1 - 2\sqrt{T}} \quad (16)$$

which is the desired relation.

# Cascade-Connected Attenuators\*

R. W. BEATTY†, MEMBER, IRE

TWO OR MORE calibrated attenuators are often connected in series (cascade), as in Fig. 1, in order to obtain a desired reduction in power level. The



Fig. 1—Cascade-connected attenuators.

total attenuation of the combination is usually obtained by adding the attenuation of each unit. In general, this will not give the correct result because of reflections due to mismatch at the attenuator junctions.

The correct value of attenuation can be obtained in terms of reflection coefficient measurements at the attenuator junctions.

For two units, the total attenuation  $A_T$  is

$$A_T = A' + A'' + 20 \log_{10} |1 - \Gamma_2' \Gamma_1''|, \quad (1)$$

where

\* Decimal classification: R396. Original manuscript received by the Institute, December 28, 1949; abstract received, May 25, 1950. This is an abstract of a paper published in *Jour. Res. Nat. Bur. Stand.*, RP2129, vol. 45; September, 1950.

† Central Radio Propagation Laboratory, National Bureau of Standards, Washington, D. C.

$A'$  = attenuation of first unit

$A''$  = attenuation of second unit

$\Gamma_2'$  = reflection coefficient at junction looking toward first unit with input matched

$\Gamma_1''$  = reflection coefficient at junction looking toward second unit with output matched.

The last term in equation (1) is the error ( $\epsilon_2$ ) resulting when the individual attenuations are added to obtain the total attenuation of the combination. If the voltage standing-wave ratios corresponding to  $\Gamma_2'$  and  $\Gamma_1''$  are measured,  $\epsilon_2$  will lie between two limits, depending upon the phases of  $\Gamma_2'$  and  $\Gamma_1''$ . A nomogram which gives the limits of error appears in Fig. 2.

If more than two cascade-connected attenuators are considered, the total attenuation may be obtained by successive use of (1), starting from one end of the chain. When the attenuation of the first two attenuators has been determined, they are treated as the first unit and the third attenuator becomes the second unit in (1). If the attenuators are numbered from the end, the following equation is obtained for any number ( $n$ ) of cascade-connected attenuators:

$$A_T = \sum_{K=1}^n A^{(K)}$$

$$+ \sum_{K=2}^n 20 \log_{10} |1 - \Gamma_2^{(K-1)} \Gamma_1^{(K)}|, \quad (2)$$

where

$(K)$  = a superscript denoting the number of the attenuator

$A^{(K)}$  = attenuation of  $K^{\text{th}}$  attenuator

$\Gamma_2^{(K-1)}$  = reflection coefficient measured at the output terminals of the  $(K-1)^{\text{th}}$  attenuator looking toward the input with the end of the chain matched

$\Gamma_1^{(K)}$  = reflection coefficient measured at the input terminals of the  $K^{\text{th}}$  attenuator with its output matched.

As shown in Fig. 1, each attenuator in the chain has its output terminals connected to the input terminals of the following attenuator.

These results may be applied to attenuators which are not used in a transmission-line system if expressions of the form  $Z - Z_0 / Z + Z_0$  are substituted for the appropriate reflection coefficients, and the same reference impedance ( $Z_0$ ) is used throughout.

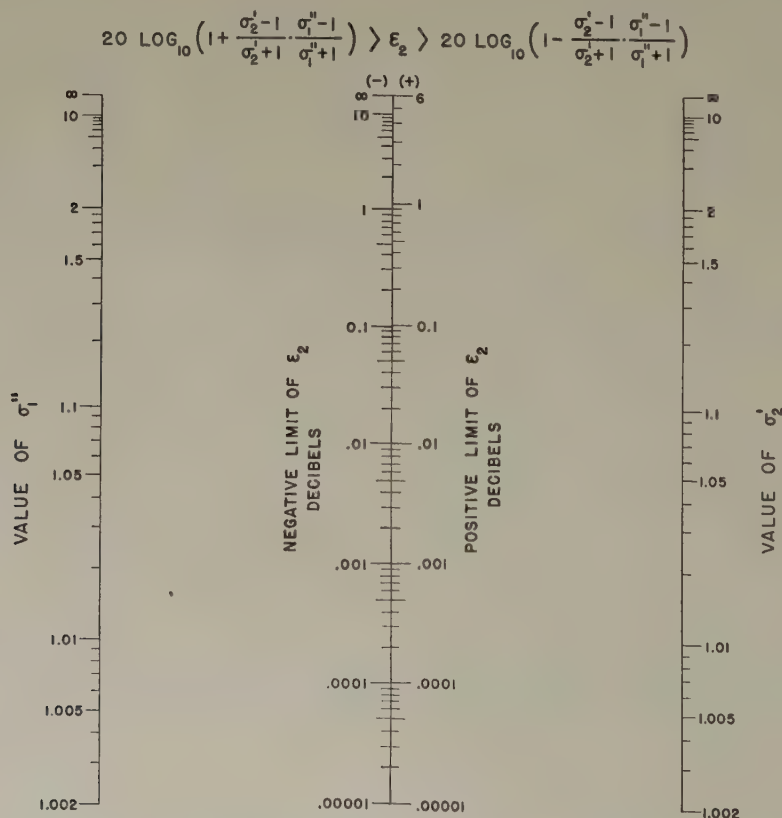


Fig. 2—Nomogram of mismatch error for two cascade-connected attenuators.

# The Design of Frequency-Compensating Matching Sections\*

V. H. RUMSEY†

**Summary**—Frequency-compensating networks have been employed to improve the impedance bandwidth of antennas and other radio-frequency components. The general problem considered here is that of transforming an impedance which changes with frequency to a specified resistance which is, as nearly as possible, constant with frequency. A simple procedure for solving the general problem is established, and formulas for the design of the appropriate matching section are worked out. The formulas give the parameters of the matching section in terms of the impedance of the load at selected frequencies. The technique is mainly applicable where the half bandwidth is small compared with the center frequency. For the purpose of illustration, the method is described in terms of its application to coaxial lines, although it is equally applicable to any form of transmission line whose characteristic impedance is known.

## INTRODUCTION

THE TECHNIQUE of compensating for the change of impedance with frequency of a load, such as an antenna, was first brought to the attention of the author in 1942, by Westcott at TRE Great Malvern, England. Techniques of frequency compensation have been widely used in the meantime, although there are relatively few open publications on the subject.<sup>1-5</sup> These consist largely of particular examples which illustrate the mechanism of frequency compensation but do not give an explicit solution of the general problem. In this paper a more general approach is attempted. A procedure and design formulas are established from which an optimum matching section can be constructed without recourse to methods of trial and error. For the purpose of illustration, the method is described in terms of its application to coaxial lines, although it is equally applicable to any form of transmission line whose characteristic impedance is known.

The problem is to design a matching section which, as nearly as possible, transforms the load impedance into

a constant resistance  $Z_0$  at all frequencies in the band. The design of the optimum matching section is based on the impedance curve obtained by plotting the variation of load impedance with frequency on the Smith Chart. While the technique of frequency compensation can be used to give considerable improvement, there is no substitute for a good impedance curve in the first place. In practice it is often possible to make adjustments to the load itself, yielding a variety of curves. The first step is therefore to select the best curve. This may be done by plotting the impedance curve with respect to a complex characteristic impedance equal to the impedance at the mean frequency. As a general rule, the best curve is the shortest curve so obtained.

## TWO-ELEMENT MATCHING SECTIONS

It is assumed that the simplest and smallest matching section is required in all cases. This means that a better match could be obtained in some cases by using a more complicated section, but it is seldom worth the trouble.

In order to set up a systematic procedure, it is convenient to restrict ourselves to quarter-wave stubs and transformers. This results in practically no loss of generality, though in some cases it may not give the most elegant solution. Fortunately these cases are uncommon in practice.

Fig. 1 shows the five types of matching section in co-

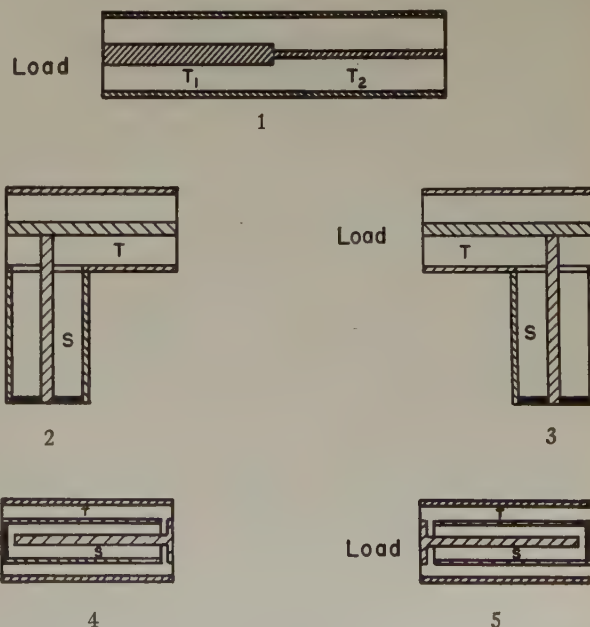


Fig. 1—Two-element frequency-compensating matching sections.

\* Decimal classification: R117.12. Original manuscript received by the Institute, July 6, 1949; revised manuscript received, April 17, 1950. Presented, 1949 IRE National Convention, New York, N. Y., March 8, 1949.

The work described in the paper was done while the author was employed by the British Air Commission, Washington, D. C., at the Naval Research Laboratory, Washington, D. C. It was originally published in report CRG 89, Naval Research Laboratory, Washington, D. C., September, 1945.

† Ohio State University, Columbus, Ohio.

<sup>1</sup> L. H. Dawson and N. M. Rust, "A wide band linear array aerial," *Jour. IEE* (London), part IIIa, vol. 93, 1946.

<sup>2</sup> E. G. Fubini and P. S. Sutro, "A wide-band transformer from an unbalanced to a balanced line," *Proc. I.R.E.*, vol. 35, p. 1153; October, 1947.

<sup>3</sup> R. G. Fellers and R. T. Weidner, "Broad-band waveguide admittance matching by use of irises," *Proc. I.R.E.*, vol. 35, p. 1080; October, 1947.

<sup>4</sup> C. H. Westcott and F. K. Goward, "The design of broadband aerial elements for 500-600 Mc ground radar," *Jour. IEE* (London), part III, vol. 96; 1949.

<sup>5</sup> Harvard Radio Research Laboratories, "Very High Frequency Techniques," McGraw-Hill Book Co., New York, N. Y., vol. I, chap. 3; 1947.

axial form although the analysis which follows applies, in general, to balanced or unbalanced lines. Each consists of two quarter-wave elements. Referring to Fig. 1,  $T_1$  and  $T_2$  denote the characteristic impedances of the two quarter-wave transformers in 1;  $S$  and  $T$  denote the characteristic impedance of the quarter-wave stubs and transformers in 2, 3, 4 and 5. It is required to determine the values of  $T_1$  and  $T_2$  or  $S$  and  $T$  which give the best match to a given resistance  $Z_0$  over the frequency band.

We first note the limitations imposed by the use of quarter-wave elements. Since the final result is required to be a pure resistance, it follows that the impedance curve applied to the beginning of the matching section must be resistive at the mean frequency and perpendicular to the line of zero reactance. In general, the curve does not satisfy these requirements, but can be made to without significantly affecting the bandwidth. Fig. 2

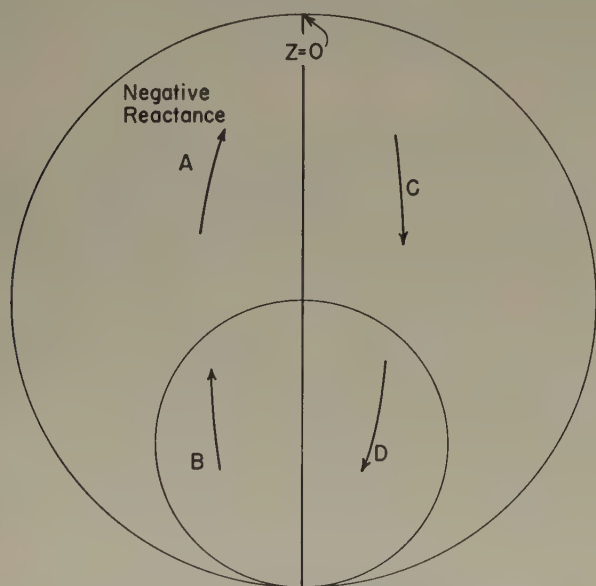


Fig. 2—Representative positions of the impedance curve on the Smith chart.

shows four typical positions of the curve. If it falls in the position  $A$  or  $B$ , then a short length of high-impedance line (effectively an inductance) will transform the midband impedance to a pure resistance,  $B$  requiring a higher impedance than  $A$ . If the curve falls in the position  $C$  or  $D$ , a short length of low-impedance line (a capacity) will transform the midband impedance to a pure resistance. In either case, since the matching section is short, the bandwidth will not be changed appreciably. We thus obtain a curve which cuts the line of zero reactance at the midband point, but not necessarily at right angles. In order to obtain the desired orientation, we must apply an appropriate length of line whose characteristic impedance is equal to the midband impedance (in order to minimize any change of bandwidth).

Hence, the compensating matching sections of Fig. 1 are preceded by two short sections, the first of which translates the curve to the real axis and the second of

which rotates it to the perpendicular position. It remains to decide which of the matching sections in Fig. 1 should be used, and to determine the values of the two characteristic impedances in it. This is accomplished by means of a series of formulas which are given below. Before giving the procedure, the method of obtaining the formulas will be described briefly. A detailed analysis is given in the Appendix.

#### DERIVATION OF THE DESIGN FORMULAS

To derive the formulas, the conditions for an exact match at the ends and the middle of the band are worked out on the assumption that the frequency band is small enough that third-order terms in the relative frequency difference (ratio of half bandwidth to mean frequency) may be ignored in comparison with first- and second-order terms. This means that the formulas cannot be used for bandwidths which exceed about 50 per cent (total). The values of, say,  $T_1$  and  $T_2$  (see Fig. 1), which give an exact match at one end of the band, also give an exact match at the other end if the curve is symmetrical about the line of zero reactance. If these values of  $T_1$  and  $T_2$  were used, the final curve would resemble curve  $A$  in Fig. 3. The nearest values of  $T_1$  and  $T_2$  which give an exact match at the midband produce a result like curve  $B$  in Fig. 3. The optimum value of, say,  $T_1$ ,

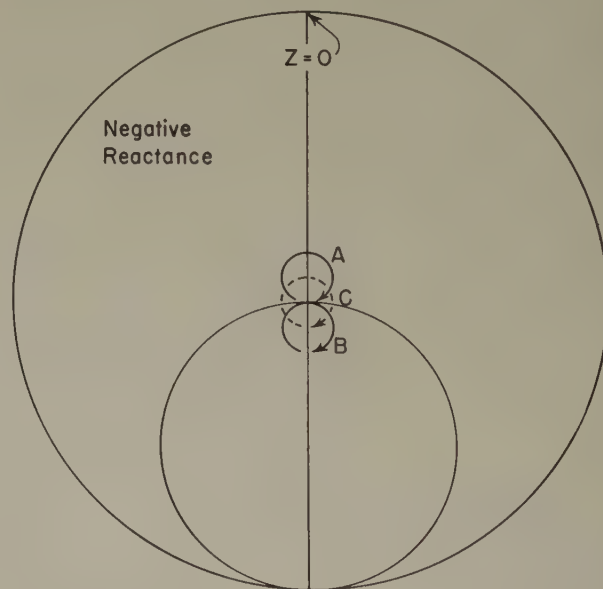


Fig. 3—Illustrating performance of optimum matching section.

is then the geometric mean of the two values which give the curves  $A$  and  $B$ ; similarly for  $T_2$ . When these mean values are used, the final curve resembles curve  $C$  in Fig. 3, for which the mismatch is roughly the same at all frequencies in the band. The value of this mismatch is easily deduced, and is given in the formulas as a voltage standing-wave ratio, which may be either greater than or less than one. The optimum values of

$S$  and  $T$  in the other matching sections are similarly determined.

The curve may not be exactly symmetrical about the zero reactance line. In order to take this into account, let  $r_L + jx_L$ ,  $r_H - jx_H$  represent the impedance points at the lowest and highest frequencies. Then in the formulas we use

$$r = \frac{1}{2}(r_L + r_H) \quad \text{and} \quad x = \frac{1}{2}(x_L + x_H)$$

and assume that  $r \pm jx$  represents the impedance at the lowest and highest frequencies; i.e., we assume that an asymmetrical curve would give about the same result as the symmetrical curve which most nearly coincides with it.

### THE DESIGN FORMULAS

The formulas apply to the frequency band between  $f - Df$  and  $f + Df$ ,  $f$  representing the mean frequency and  $Df$  the half bandwidth. The values of the characteristic impedances in the matching sections are given relative to  $Z_0$ , the resistance to which a match is required. The numbers 1, 2, 3, 4, and 5, associated with the formulas, apply to the corresponding matching sections given in Fig. 1.

Frequency	Impedance $\div Z_0$	Admittance $\times Z_0$
$f - Df$	$r_L + jx_L$	$g_L + jb_L$
$f$	$r_1$	$g_1$
$f + Df$	$r_H - jx_H$	$g_H - jb_H$

$$d = \tan \frac{\pi Df}{2f} \quad M = \text{VSWR mismatch}$$

$$2r = r_H + r_L \quad 2g = g_H + g_L$$

$$2x = x_H + x_L \quad 2b = b_H + b_L$$

1. For  $x > 0$

$$h = \sqrt{r} - d^2(r + \sqrt{r})$$

$$M = \frac{\sqrt{r_1}}{h}$$

$$a = \frac{x(1 - d^2h)}{2dh(1 + h)}$$

$$t_2 = \frac{h^{1/4}}{r_1^{1/8}} \left( a + \sqrt{a^2 + \frac{r}{h}} \right)$$

$$t_1 = t_2 \sqrt{h \sqrt{r_1}}$$

1. For  $b > 0$ , replace  $r_1$ ,  $r$ ,  $x$ ,  $t_1$ , and  $t_2$  by  $g_1$ ,  $g$ ,  $b$ ,  $1/t_1$ , and  $1/t_2$ , respectively.

$$2. M = \sqrt{\frac{g}{g_1}}$$

$$t^2 = \frac{1}{\sqrt{gg_1}}$$

$$\frac{1}{s} = \sqrt{g}(g - 1) + \frac{b}{d}$$

$$3. M = \frac{\sqrt{r} - dx}{\sqrt{r_1}}$$

$$t^2 = \sqrt{r_1}(\sqrt{r} - dx)$$

$$\frac{1}{s} = \frac{1}{\sqrt{r}} \left[ r - 1 + \frac{x}{dt} \right]$$

4. Replace  $r$ ,  $r_1$ ,  $x$ ,  $s$ , and  $t$  in 3 by  $g$ ,  $g_1$ ,  $b$ ,  $1/s$ , and  $1/t$ , respectively.

5. Replace  $g$ ,  $g_1$ ,  $b$ ,  $1/s$ , and  $1/t$  in 2 by  $r$ ,  $r_1$ ,  $x$ ,  $s$ , and  $t$ , respectively.

### PROCEDURE

Summarizing the foregoing discussion, we obtain the following procedure for matching a given load over a band of frequencies  $f \pm Df$  to a resistance  $Z_0$ :

1. All possible adjustments are made to load in order to obtain the best curve in the first place.

2. The curve is transformed using methods which do not reduce the bandwidth until

- (a) the midband impedance is a pure resistance,
- (b) the curve cuts the line of zero reactance at right angles.

3. The curve so obtained is plotted with respect to  $Z_0$  and values of the parameters required in the formulas are noted.

4. The best matching section is determined by substitution in the formulas for mismatch.

5. The design of the best matching section is determined from the formulas.

### AN EXAMPLE

Suppose that the best curve obtained was curve  $A$  in Fig. 4, plotted with respect to  $Z_0$  over a  $\pm 10$  per cent

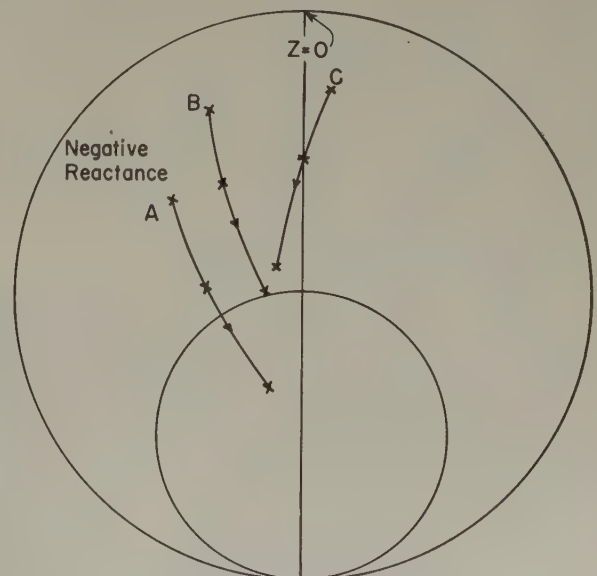


Fig. 4—Transformation of initial curve  $A$  to real axis. (Frequency increases in direction of arrow.)

band. Referring to the procedure, step 2(a) is accomplished by a short length of high-impedance line. Suppose that  $2Z_0$  is the highest convenient value. Replotting  $A$  with respect to  $2Z_0$  we obtain curve  $B$ . In order to make the impedance at the mean frequency  $f$  a pure resistance, it is necessary to transform  $B$  through a length of  $0.05\lambda$  (at frequency  $f$ ), of characteristic impedance  $2Z_0$ . After this transformation we obtain curve  $C$  in Fig. 4, which gives an impedance at frequency  $f$  of  $0.7Z_0$ . To accomplish step 2(b), we therefore transform through an appropriate length of line of characteristic impedance equal to  $0.7Z_0$ . Replotting curve  $C$  in Fig. 4 with respect to  $0.7Z_0$  gives curve  $D$  in Fig. 5. Curve  $D$  shows

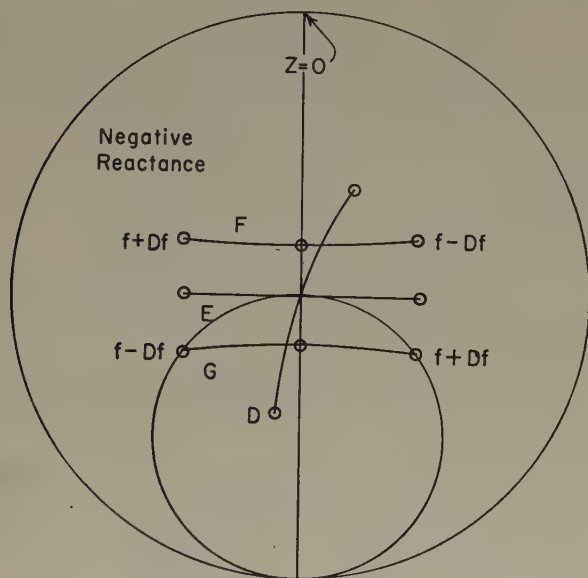


Fig. 5—Transformation of curve  $C$ , Fig. 4, to orientation perpendicular to real axis.

that a line of length  $0.1\lambda$  (at frequency  $f$ ) of characteristic impedance  $0.7Z_0$  transforms the curve so that it cuts the line of zero reactance at right angles, giving curve  $E$  in Fig. 5. Proceeding to step 3, curve  $E$  is plotted with respect to  $Z_0$ , giving curve  $F$ , from which we plot the admittance curve  $G$  in Fig. 5. Curves  $F$  and  $G$  are the impedance and admittance curves which are used in the formulas. Hence, before using one of the matching sections shown in Fig. 1, we apply  $0.05\lambda$  of  $2Z_0$  line followed by  $0.1\lambda$  of  $0.7Z_0$  line.

The values of  $r_L$ ,  $r_H$ ,  $x_L$ ,  $x_H$ , and  $r_1$  are read from curve  $F$ , Fig. 5, and the values of  $g_L$ ,  $g_H$ ,  $b_L$ ,  $b_H$ , and  $g_1$  are read from curve  $G$ , Fig. 5. Applying step 4 to formulas 1, 3, and 5 we obtain 1.2, 1.33, and 1.18, respectively, for the mismatch (when expressed as a fraction greater than 1). Matching sections 2 and 5 give a negative value of  $S$  and are therefore not applicable. Using the formulas, we obtain the values  $T=0.77Z_0$  and  $S=2.77Z_0$  as the characteristic impedances of the elements in the matching section 5. The complete matching section constructed in coaxial line thus takes the form shown in Fig. 6.

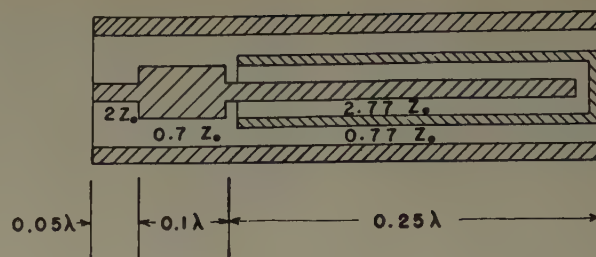


Fig. 6—Coaxial-line form of optimum matching section for curve  $A$ , Fig. 4.

In order to illustrate the action of the matching sections, the transformations have been plotted with respect to  $Z_0$  for matching section 5 in Fig. 7. The initial curve is represented by  $A$ , curve  $B$  represents the impedance after the addition of the series stub, and curve  $C$  represents the final impedance. Curve  $C$  in Fig. 7 is slightly overcompensated, indicating that a smaller value of  $S$  could have been used in matching section 5.

It should be remembered that there is an inevitable error in any impedance measurement. Also, that a change of physical dimensions in a transforming section gives rise to a small reflection, which is not taken into account in the formulas. In general, the accuracy of the formulas given here is about the same as the accuracy of measurement.

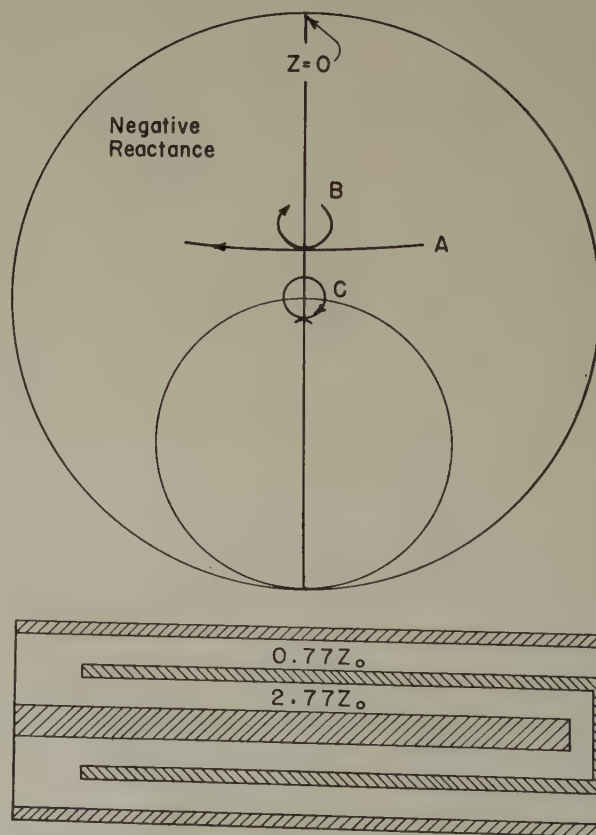


Fig. 7—Transformations due to frequency-compensating matching section.

## APPENDIX

*The Derivation of the Formulas*

Referring to Fig. 1, let the load be connected to the transformer of characteristic impedance  $T_1$  in matching section 1. The impedance of the load is  $R+jX$  at the lowest frequency,  $R_1$  at the mean frequency, and  $R-jX$  at the highest frequency.

Let

$$d = \tan \frac{Df}{f} 90^\circ \quad (1)$$

where  $f \pm Df$  are the highest and lowest frequencies.

Let

$$z = \frac{Z}{Z_0}, \quad r = \frac{R}{Z_0}, \quad x = \frac{X}{Z_0}, \quad t_1 = \frac{T_1}{Z_0}, \quad t_2 = \frac{T_2}{Z_0}, \text{ and so forth,}$$

where  $Z_0$  is the characteristic impedance to which a match is required. Then the impedance seen through the first quarter-wave transformer of characteristic impedance  $T_1$  at the frequency  $f-Df$  is given by  $Z'$  where

$$\frac{z'}{t_1} = \frac{zd + jt_1}{t_1d + jz} \quad (2)$$

Similarly the final impedance is given by

$$\frac{z''}{t_2} = \frac{z'd + jt_2}{t_2d + jz'} \quad (3)$$

The condition that  $Z''=Z_0$  or  $z''=1$  results in the equations

$$t_1^2 - rt_2^2 + dx(t_1 + t_2) + d^2t_1t_2(r-1) = 0 \quad (4)$$

and

$$t_2^2x + d(t_1 + t_2)(r - t_1t_2) - d^2t_1t_2x = 0. \quad (5)$$

Let

$$h = \frac{t_1}{t_2} \quad (6)$$

Then from (4) we have

$$t_2 = \frac{-dx(1+h)}{h^2 - r + d^2h(r-1)} \quad (7)$$

Elimination of  $t_1$  and  $t_2$  from (5), (6), and (7) gives

$$(h^2 - r)(u^2 - g) + 2d^2[gh(1-r) + ru(1-g) + 1 - rg] + d^4(r-1)(g-1) = 0 \quad (8)$$

where

$$uh = 1 \quad (9)$$

$$g = GZ_0 = \frac{r}{r^2 + x^2} = \text{normalized conductance.} \quad (10)$$

Equation (8) is unaltered if we interchange  $h$  and  $u$ , and  $r$  and  $g$ . This means that if we find a solution in terms of

impedances we may immediately write down an alternative solution in terms of admittances by making the above interchange.

We now make use of the fact that  $d$  is small. For a  $\pm 10$ -per cent band,  $d=0.16$ ,  $d^2=0.0256$ , and so forth. An approximate solution is therefore obtained by assuming  $d=0$ . This gives

$$h^2 = r \quad \text{or} \quad u^2 = g,$$

i.e.,

$$h^2 = r \quad \text{or} \quad h^2 = \frac{1}{g} = \frac{r^2 + x^2}{r} \quad (11)$$

Now these are two different solutions if  $x \neq 0$ . To determine which is correct we substitute in (7).

Suppose  $x > 0$  and  $h^2 = r$ .

Then

$$t_2 = \frac{\text{negative quantity}}{d^2h(r-1)} \quad (12)$$

Therefore  $t_2$  exists only if  $r < 1$

Suppose  $x > 0$  and

$$h^2 = r + \frac{x^2}{r}.$$

Then

$$t_2 = \frac{\text{negative quantity}}{\frac{x^2}{r} + d^2h(r-1)}$$

Therefore,  $t_2$  exists only if

$$d^2h(1-r) > \frac{x^2}{r} \quad (13)$$

Since we have assumed  $x \neq 0$ , but  $d=0$ , the inequality (13) is plainly inadmissible. The inequality (12) which restricts the values of  $r$  arises because the original approximation was poor, not because the assumption that  $h^2=r$  is fundamentally incorrect.

Suppose  $x < 0$  and  $h^2 = r$ .

Then

$$t_2 = \frac{\text{positive quantity}}{d^2h(r-1)}.$$

Therefore,  $t_2$  exists only if  $r > 1$ .

Suppose

$$x < 0 \quad \text{and} \quad h^2 = r + \frac{x^2}{r}.$$

Then

$$t_2 = \frac{\text{positive quantity}}{\frac{x^2}{r} + d^2h(r-1)}.$$

Therefore,  $t_2$  exists only if

$$\frac{x^2}{r} > d^2 h(1 - r). \quad (15)$$

Since  $d$  is small, the inequality (15) is less restrictive than the inequality (14). We conclude therefore that the correct solution to the first approximation is

$$h^2 = r \text{ if } x > 0 \text{ and } u^2 = g \text{ if } x < 0.$$

To proceed with the solution, we return to (8). Assuming  $x > 0$ , we substitute the first approximation  $h^2 = r$  in the coefficient of  $d^2$ , ignoring  $d^4$ , giving

$$(h^2 - r)(u^2 - g) = 2d^2(\sqrt{r} + 1)(gr - 1).$$

Solving for  $h$  gives to the same order of approximation

$$h = \sqrt{r} - d^2(r + \sqrt{r}). \quad (16)$$

This approximation is sufficiently accurate for most practical purposes, since the next approximation involves terms in  $d^4$  which is very small.

We can now solve for  $t_2$  from either (4) or (5). Equation (4) gives a poor approximation when  $x$  is small. We therefore use (5), which may be written

$$t_2^2 - 2at_2 - \frac{r}{h} = 0 \quad (17)$$

where

$$a = \frac{x(1 - d^2 h)}{2dh(1 + h)}. \quad (18)$$

Therefore, when  $x > 0$ ,

$$t_2 = a + \sqrt{a^2 + \frac{r}{h}}. \quad (19)$$

We do not consider the negative square root because this would make  $t_2$  negative.  $t_1$  is then given by

$$t_1 = ht_2. \quad (20)$$

If the impedance at the highest frequency is  $R - jX$ , we obtain identical equations. Hence, the conditions for an exact match at the ends of the band are identical and are given by the above equations. The condition for an exact match at the midband frequency is simply

$$t_1 = t_2 \sqrt{r_1} \quad (21)$$

where  $R_1 + j0$  is the impedance at the midband frequency.

Summarizing the results, the condition for a match at the ends of the band when  $x > 0$  is

$$t_1 = i \quad (22)$$

$$t_2 = k \quad (23)$$

where  $i$  and  $k$  are given by (20) and (19).

The condition for a perfect match at the middle of the band is

$$\frac{t_1}{t_2} = L \quad (24)$$

where  $L$  is given by (21).

The values of  $t_1$  and  $t_2$  which satisfy (24) and deviate least from those given by (22) and (23) are given by

$$\frac{t_1}{t_2} = L \quad t_1 t_2 = ik \quad (25)$$

whence

$$t_1^2 = Lik \quad (26)$$

$$t_2^2 = \frac{ik}{L} \quad (27)$$

are the values of  $t_1$  and  $t_2$  which give an exact match at the midband and have the smallest relative deviation from the values which give an exact match at the ends of the band. The values of  $t_1$  and  $t_2$  which most nearly satisfy (22), (23), (26), and (27) are given by

$$t_1^2 = i\sqrt{Lik} = i^{3/2}k^{1/2}L^{1/2} \quad (28)$$

$$t_2^2 = k\sqrt{\frac{ik}{L}} = i^{1/2}k^{3/2}L^{-1/2}. \quad (29)$$

Substituting the values of  $i$ ,  $k$ , and  $L$  we have

$$t_1 = h^{3/4} \left[ a + \sqrt{a^2 + \frac{r}{h}} \right] r_1^{1/8} \quad (30)$$

$$t_2 = h^{1/4} \left[ a + \sqrt{a^2 + \frac{r}{h}} \right] r_1^{-1/8}. \quad (31)$$

When  $t_1$  and  $t_2$  are given by (30) and (31) the mismatch is very nearly the same at all frequencies. We may therefore find its value by evaluating at the midband. The impedance obtained from (30) and (31) at the midband is

$$N = R_1 \frac{t_2^2}{t_1^2} = \frac{R_1}{h\sqrt{r_1}}. \quad (32)$$

This represents a mismatch  $M$  on a line of characteristic impedance  $Z_0$  where

$$M = \frac{N}{Z_0} = \frac{R_1}{Z_0}, \quad \frac{1}{h\sqrt{r_1}} = \frac{\sqrt{r_1}}{h} = \frac{r_1 t_2^2}{t_1^2}. \quad (33)$$

Hence the complete solution for  $x > 0$  is given by (16), (18), (30), (31), and (33). For  $x < 0$  the solution is obtained by interchanging impedance and admittance.

The formulas for the other matching sections are similarly derived.

# Signal-to-Noise Improvement Through Integration in a Storage Tube\*

J. V. HARRINGTON†, ASSOCIATE, IRE, AND T. F. ROGERS†, MEMBER, IRE

**Summary**—Random noise places a fundamental physical limitation on the precision with which a signal may be observed or measured. In the case where signal energy arrives piecemeal in a repetitive manner, this energy may be integrated over a period of time to provide an increase in signal-to-noise ratio, thereby reducing this limitation. Several techniques have been suggested for performing this integration process; recent work with an electronic barrier-grid storage tube has shown that such a device exhibits many desirable features when used as an integrating device. An introduction to integrator theory in general and an analysis of the storage tube operating as an integrator are given. Calculations are made of the expected improvement in signal-to-noise ratio as a function of tube parameters and number of integrations. Evidence is presented to show that the improvement achieved experimentally is nearly equal to the calculated value.

## I. INTRODUCTION

THE BENEFITS to be obtained by applying an integration, or addition, process to periodic signals in noise have been appreciated for some time. Various practical integrators have been tried in attempts to obtain some of these benefits, since, especially in the field of radar, where signals are essentially repetitive over a number of intervals, an improvement in signal-to-noise ratio represents an improvement in the entire target detection process.<sup>1</sup> Long-persistence cathode-ray tubes and photographic integration techniques have been used; narrow band-pass filters constructed around the pulse-repetition frequency have also come into prominence; delay lines also are now being used. Each of these methods, while possessing certain advantages, also presents more or less distinct limitations.

Regardless of the method used in any particular integration scheme, a fundamental requirement which they all have in common is that of a suitable "memory." This memory must possess the faculty of accepting and remembering with reasonable accuracy a number of mixed signals, that is, a mixture of signals and the contaminating noise. When a number of such sequences have been added in storage, their sum may be extracted and examined. With the knowledge that the noise, in general, will build up at a slower rate than will the coherent signal, an increasingly better judgment may be made on the composite signal as to which is the signal and which is the noise.

\* Decimal classification: R361.211×R339. Original manuscript received by the Institute, December 8, 1949; revised manuscript received, June 2, 1950. Presented, 1949 IRE National Convention, New York, N. Y., March 10, 1949.

† Air Force Cambridge Research Laboratories, Cambridge, Mass.

<sup>1</sup> S. Goldman, "Some Fundamental Considerations Concerning Noise Reduction and Range in Radar and Communications," M.I.T. Technical Report No. 32, Dec. 15, 1947. Also published, *Proc. I.R.E.*, vol. 36, pp. 584-594; May, 1948.

Certain types of storage tubes possess characteristics which indicate that they might serve as memory and general comparative devices. The STE-A barrier grid storage tube<sup>2</sup> is one such device and this paper will deal with its theory and performance as a practical signal integrator, as well as with the theory of signal-to-noise ratio improvement in general.

## II. THEORETICAL ANALYSIS OF POST-DETECTOR INTEGRATION

While there are many factors which limit the benefits to be obtained by the addition or integration of coherent video signals in noise, the two which appear to predominate in importance are (1) the small signal suppression effect attributable to the inherent characteristics of the second detector; and (2) the effect of a nonideal integrator law, i.e., one in which the component signals are not equally weighted in forming the sum or total integrated signal.

The first of these effects has been discussed many times as part of the general problem of analyzing the passage of signal-plus-noise through a nonlinear circuit. This general problem has been attacked by several authors but, in the main, they have been concerned with calculating the power spectrum of the detector output. As Rice<sup>3</sup> has pointed out, there also exists the question of determining the probability distribution function in the output of a nonlinear device given the statistical nature of the input. It is with this aspect of the problem that we shall be concerned, since the probability distribution is, in general, of greater interest in analyzing the effect of integration on the detected mixed signal.

Specifically, this analysis will deal with the calculation of the resultant distribution function where a sine wave (carrier) plus noise is (a) processed by a nonlinear detector, and (b) further processed by a nonideal integrator. The usual assumptions are made: that the noise in the intermediate-frequency amplifier preceding the second detector is normally distributed, that the power spectrum of the noise is confined to a small band compared to the carrier frequency, and that the detector (and video amplifier) filter doesn't seriously affect the spectrum of the low-frequency portion of the detected mixed signal. Under these conditions the output of the

<sup>2</sup> A. S. Jensen, J. P. Smith, M. H. Mesner, and L. E. Flory, "Barrier grid storage tube and its operation," *RCA Rev.*, vol. 9, p. 3; March, 1948.

<sup>3</sup> S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. Jour.*, vol. 25, p. 151; January, 1945.

detector may be obtained from the envelope of the mixed signal at the input and, in fact, the detector may then be alternately described as "an envelope tracer" as suggested by North.<sup>4</sup> The envelope of a sine wave plus noise has the probability distribution

$$P(R)dR = \frac{R}{\psi_0} e^{-(R^2 + P^2/2\psi_0)} I_0\left(\frac{RP}{\psi_0}\right) dR. \quad (1)$$

The moments of this distribution are given by Rice<sup>8</sup> as

$$\overline{R^n} = (2\psi_0)^{n/2} \Gamma(1 + n/2) {}_1F_1(-n/2; 1; -P^2/2\psi_0). \quad (2)$$

The notation is made somewhat simpler if we normalize with respect to the root-mean-square noise by letting

$$v = R(\psi_0)^{-1/2}; \quad a = P(\psi_0)^{-1/2}; \quad x = a^2/2. \quad (3)$$

Then

$$\overline{v^n} = 2^{n/2} \Gamma(1 + n/2) {}_1F_1(-n/2; 1; -x). \quad (4)$$

For a half-wave  $k$ th law detector the normalized output amplitude  $Z$  is defined as

$$Z = Bv^k \quad \text{for } v > 0. \quad (5)$$

For this type of detector, which is perhaps the most common of all, (that is, neither biased nor saturated) the output distribution is very simply described in terms of its moments since

$$\overline{Z^n} = B^n \overline{v^{nk}}. \quad (6)$$

Hence, the  $n$ th moment of the distribution of the output of a  $k$ th law detector is the same as the  $kn$ th moment of the output of a linear detector and is given by

$$\overline{Z^n} = B^n 2^{nk/2} \Gamma(1 + nk/2) {}_1F_1(-nk/2; 1; -x). \quad (7)$$

The distribution function  $p(Z)$  may then be expressed as an Edgeworth<sup>5</sup> series since the various coefficients involved in the series are defined by the moments. The predominant terms in this series are

$$p(y)dy = \left[ \phi(y) - \frac{\alpha_3}{3!} \phi^3(y) + \frac{(\alpha_4 - 3)}{4!} \phi^4(y) + \frac{10}{6!} \alpha_3^2 \phi^6(y) \right] dy \quad (8)$$

where

$$y = \frac{Z - m}{\sigma}; \quad m = \overline{Z}; \quad \sigma^2 = \overline{Z^2} - m^2, \\ \phi(y) = (2\pi)^{-1/2} e^{-y^2/2},$$

and

$$\phi^n(y) = (d^n/dy^n) \phi(y) = (-1)^n (2\pi)^{-1/2} e^{-y^2/2} H_n(y).$$

<sup>4</sup> D. O. North, "An Analysis of the Factors Which Determine Signal-Noise Discrimination in Pulsed Carrier Systems," RCA Laboratories Report PTR-6-C, June, 1943.

<sup>5</sup> H. Cramer, "Mathematical Methods of Statistics," Princeton University Press, Princeton, N. J., p. 229; 1946.

The factors  $\alpha_3$  and  $\alpha_4$  are, respectively, the third and fourth standard central moments and are measures of the skewness and peakedness of the distribution or, more exactly, one can define<sup>6</sup>

$$\begin{aligned} \zeta_3 &= \alpha_3 = \mu_3/\sigma^3 \\ &= \text{coefficient of skewness} \\ \zeta_4 &= [\alpha_4 - 3] = \mu_4/\sigma^4 - 3 \\ &= \text{coefficient of excess or kurtosis.} \end{aligned} \quad (9)$$

If now  $n$  signals each having the amplitude distribution given by (8) are added with the same weight being given to each signal, it may be shown that

$$p_n(y)dy = dy \left[ \phi(y) - \frac{\zeta_3}{3!\sqrt{n}} \phi^3(y) + \frac{\zeta_4}{4!n} \phi^4(y) + \dots \right] \quad (10)$$

where, now,  $y = (Z - nm)/n^{-1/2}\sigma$ , and we observe that one effect of integration is to diminish the skewness and peakedness of the distribution, forcing it to approach a normal distribution specified by the number  $n$  and the first two moments of the original distribution. This leads us to the definition of signal-to-noise ratio as the mean value of the mixed signal divided by its root-mean-square value where the reference axis chosen in specifying this mean is the average value of the detected noise alone. The reason for the shift of axis can be seen from Fig. 1. A similar definition of signal-to-noise ratio has

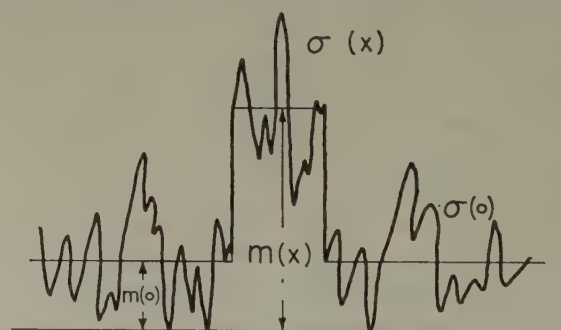


Fig. 1—The shift in reference axis for low signal-to-noise ratios.

been used by Van Vleck and Middleton.<sup>6</sup> In terms of the moments previously calculated (8), the signal-to-noise ratio at the output of a  $k$ th law detector is therefore

$$\begin{aligned} \frac{\overline{m}}{\sigma_x} &= \frac{m(x) - m(0)}{\sigma(x)} \\ &= \frac{{}_1F_1(-k/2; 1; -x) - 1}{\left[ \frac{\Gamma(k+1)}{\Gamma^2(1+k/2)} {}_1F_1(-k; 1; -x) - {}_1F_1^2(-k/2; 1; -x) \right]^{1/2}} \end{aligned} \quad (11)$$

<sup>6</sup> J. H. Van Vleck and D. Middleton, "A theoretical comparison of the visual, aural and meter reception of pulsed signals in the presence of noise," *Jour. Appl. Phys.*, vol. 17, p. 940; November, 1946.

While this is a rather cumbersome expression to analyze, it is possible to make approximations for very large and very small values of  $x$  from which the consequences of small signal detection can be deduced. Thus, for  $x \ll 1$  we take the first two terms of the series expansion for the confluent hypergeometric function

$${}_1F_1(-k; 1; -x) \cong (1 + kx). \quad (12)$$

Using this, for small signal-to-noise ratios (11) becomes

$$\bar{m}/\sigma_x \cong Kx = Ka^2/2 \quad (13)$$

where

$$K = (k/2)([\Gamma(k+1)/\Gamma^2(1+k/2)] - 1)^{-1/2}.$$

The factor  $K$  is a function only of the power of the detector law and may be considered a sort of dynamic amplification factor which has the maximum value unity for the square law detector and falls off for all others; its reciprocal has been calculated by North in some unpublished work.<sup>4</sup> From (13) one draws the now well-known conclusion<sup>7</sup> that for detection of small signals, all detectors (except for the scale factor  $K$ ) act like the square-law detector. If a linear integrator is used to bring the signal-to-noise ratio up to a given level by  $n$  additions, then the output signal-to-noise ratio in terms of the input  $S/N$  ratio is given by

$$n^{1/2}\bar{m}/\sigma_x = K(n)^{1/2}a^2/2 = (K/2)(n^{1/4}a)^2. \quad (14)$$

A rather serious consequence of this small signal suppression effect is to cause the  $S/N$  improvement factor to go up as the  $1/4$  power of the number of additions for  $x \ll 1$  instead of the  $1/2$ -power law predictable for predetector integration.<sup>8</sup> (In predetector integration,  $S/N \cong a$ .)

For very large input signal-to-noise ratios ( $x \ll 1$ ) we use the first few terms of the asymptotic expansion<sup>3</sup>

$${}_1F_1(-k; 1; -x) \cong \frac{x^k}{\Gamma(k+1)} \left[ 1 + \frac{k^2}{1!x} + \frac{k^2(k-1)^2}{2!x^2} + \dots \right]. \quad (15)$$

This leads to the relationship

$$\bar{m}/\sigma_x \cong k(2x)^{1/2} = a/k; \quad \text{for } x \gg 1. \quad (16)$$

Thus, for large signal-to-noise ratios, a linear relationship exists between the input and detector output which yields an integration improvement factor varying as the half power of  $n$ . In between these two limiting cases for very small and very large  $x$  one would expect a transition region to exist in which the improvement factor varies smoothly from a  $1/2$ -power law to a  $1/4$ -power law going from large to small input signal-to-noise ratios.

<sup>7</sup> D. Middleton, "Rectification of a sinusoidally modulated carrier in the presence of noise," *Proc. I.R.E.*, vol. 36, p. 1467; December, 1948.

<sup>8</sup> See page 24 of footnote reference 1.

### III. PRACTICAL INTEGRATORS

The discussion thus far has been based on the existence of a perfect integrator in which the weight a signal is given in forming the sum or total signal is independent of its time of arrival. In general, such devices are not practically realizable; consequently, our purpose now is to analyze the operation of a nonideal integrator, and to see how the results of the previous section are to be modified by the actual integrator law. In this regard it is instructive to note the general law by which signals are added in some simple integration networks prior to consideration of storage-tube integration. It may be shown, for instance, that if a succession of impulses are applied at intervals ( $\Delta t$ ) to a single  $RC$  network, then the response at the time the  $n$ th signal is applied is

$$h(t) = [1/RC][E_n + E_{n-1}e^{-\Delta t/RC} + \dots + E_1e^{-(n-1)(\Delta t/RC)}]. \quad (17)$$

The law of addition here is evidently a "weighted linear one" in which the effect of each signal is exponentially weighted. Also, the narrow-band  $RLC$  circuit, whose mechanical analogue has been used very successfully in extracting weak signals from noise,<sup>9</sup> has a resultant response after  $n$  additions of

$$h(t) = [1/C][E_n + E_{n-1}e^{-(L/C)^{1/2}\pi/R} + \dots + E_1e^{-(n-1)(L/C)^{1/2}\pi/R}] \quad (18)$$

which again involves exponential weighting. Again, integration networks, for periodic signals, may also be designed around delay lines where the delay is made equal to the repetition interval of the signal. Here, a feedback loop exists and, for stability, the loop gain must be less than unity; the summation again follows the law

$$E_r = [E_n + E_{n-1}e^{-\alpha} + \dots + E_1e^{-(n-1)\alpha}] \quad (19)$$

where  $\alpha$  is the attenuation around the loop in nepers and is greater than zero.

The STE-A barrier grid storage tube,<sup>2</sup> the integrating device to be considered here, consists of an electron gun and electrostatic beam deflection system, a dielectric surface on which the signal is stored as a charge distribution, a backing plate to which the input signal is applied, and a secondary beam collector from which the reading signal is obtained (see Fig. 2). In operation, the dielectric surface will be charged in the absence of a signal to some potential at which the effective secondary-emission ratio is unity and equilibrium exists. Thereafter, when a signal is applied to the back plate the emission ratio is changed, resulting in a charge, which is different from the equilibrium charge, being deposited on the dielectric at the point of beam impingement. Thus, for a linear beam motion, a signal varying in time would be

<sup>9</sup> F. R. Dickey, A. G. Emslie, and H. Stockman, "Extraction of Weak Signals from Noise by Integration," Cambridge Research Laboratories Report, E5038, September, 1948.

stored as a charge distribution varying in distance along a line of the dielectric surface. In reading, when no signal is applied to the back plate, the beam in sweeping over this same line acts to restore the charge distribution to the equilibrium value and, in so doing, the secondary-emission current as observed at the collector will regenerate essentially a counterpart of the stored signal.

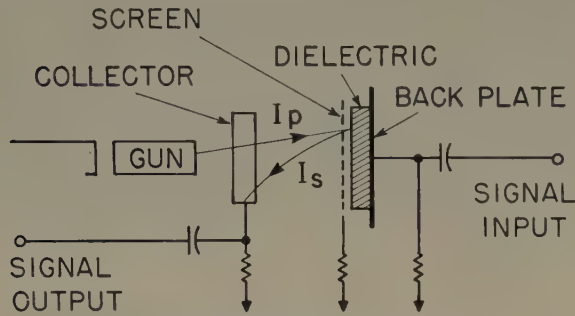


Fig. 2—Simplified schematic of barrier-grid storage tube.

While such a device is difficult to analyze in a rigorous manner, it is possible to derive certain very useful relationships descriptive of the tube's external behavior.<sup>10</sup> Such an approximate analysis, incidentally, agrees reasonably well with observed tube performance and hence, to some degree we are justified in using its results to further our present analysis. It may be shown that for storage of a succession of pulses of amplitude  $E_1, E_2, \dots, E_n$ , the maximum amplitude of the corresponding voltage distribution built up along a scan line of the dielectric surface is

$$E_r = (1 - \epsilon^{-\gamma}) [E_n + E_{n-1}\epsilon^{-\gamma} + \dots + E_1\epsilon^{-(n-1)\gamma}] \quad (20)$$

where  $\gamma = [KI_b/CV]$  is a factor of basic importance in describing tube operation.<sup>11</sup> A reading operation following this series of writing operations would give an output signal resembling the net charge distribution along the scan line, and would leave a resulting charge distribution, when the reading operation is completed, given by

$$E_r = \epsilon^{-\gamma}(1 - \epsilon^{-\gamma}) [E_n + E_{n-1}\epsilon^{-\gamma} + \dots + E_1\epsilon^{-(n-1)\gamma}]. \quad (21)$$

Thus, the reading operation is also an erasing operation in which the stored signal is reduced by an amount  $\epsilon^{-\gamma}$ .

It may be noted that many of the basic integrating devices, the RC network, narrow-band filter, and regenerative delay-line loop, all have the same general law of addition that the storage tube possesses, i.e.,

$$E_{out} = \sum_{q=1}^n E_q \epsilon^{-(n-q)\gamma}. \quad (22)$$

<sup>10</sup> J. V. Harrington, "Storage of small signals on a dielectric surface," to be published in *Jour. Appl. Phys.*, October, 1950.

<sup>11</sup> A quantity having similar significance and called the tube-discharge factor by Jensen, et al (see footnote reference 2) is equal to  $(1 - \epsilon^{-\gamma})$ .

It is of primary interest to calculate the output distribution when  $n$  mixed signals are added in accordance with this law<sup>12</sup> and not the ideal linear law previously assumed (10). If  $n$  independent variables are to be added, it may be shown that the semi-invariants of their sum are simply given by

$$X_k' = \sum_{n=1}^n X_{k_n} \quad (23)$$

where the first few semi-invariants in terms of the central moments are

$$\begin{aligned} X_1 &= n & X_3 &= \mu_3 \\ X_2 &= \sigma^2 & X_4 &= \mu_4 - 3\sigma^4. \end{aligned}$$

For addition in accordance with (22) where the scale factors<sup>13</sup> are constantly reduced in powers of  $\epsilon^{-\gamma}$  we obtain

$$\begin{aligned} X_k' &= X_k [1 + \epsilon^{-k\gamma} + \dots + \epsilon^{-k(n-1)\gamma}] \\ &= X_k [(1 - \epsilon^{-kn\gamma}) / (1 - \epsilon^{-k\gamma})]. \end{aligned} \quad (24)$$

The coefficients of the Edgeworth series may be expressed in terms of these semi-invariants normalized with respect to  $X_2$ . Thus the resulting signal-to-noise ratio, coefficient of skewness, and coefficient of excess are given by

$$\begin{aligned} \frac{m'}{\sigma} &= \frac{X_1'}{X_2'^{1/2}} = \rho_1(n, \gamma) \cdot \frac{m}{\sigma} \\ \xi_3' &= \frac{X_3'}{X_2'^{3/2}} = \rho_3(n, \gamma) \cdot \xi_3 \\ \xi_4' &= \frac{X_4'}{X_2'^2} = \rho_4(n, \gamma) \cdot \xi_4. \end{aligned} \quad (25)$$

From (24) and (25), it is seen that the general form of the improvement factor  $\rho$  is

$$\rho_k(n, \gamma) = \left[ \frac{\sinh n\gamma k/2}{\sinh \gamma k/2} \right] \left[ \frac{\sinh \gamma}{\sinh n\gamma} \right]^{k/2}. \quad (26)$$

For the first few values of  $\rho$ , and the practical assumption of  $\gamma \ll 1$ , this reduces to the more useful forms

$$\begin{aligned} \rho_1(n, \gamma) &= (n)^{1/2} \left[ \frac{\tanh n\gamma/2}{n\gamma/2} \right]^{1/2} \\ \rho_3(n, \gamma) &\cong (n)^{-1/2} (n\gamma / \tanh n\gamma)^{1/2} \\ \rho_4(n, \gamma) &= (1/n) (n\gamma / \tanh n\gamma). \end{aligned} \quad (27)$$

We conclude from this that to describe properly the effects of integration we must define not only a signal-to-noise improvement factor, but also reduction factors representing the reduction in the higher-order coefficients of the Edgeworth series. Thus,  $\rho_1$  can be termed the signal-to-noise improvement factor and  $\rho_3$  and  $\rho_4$

<sup>12</sup> It should be noted that a treatment of integrators with poor memories first appeared in the work of North. See footnote reference 4.

<sup>13</sup> See page 187 of footnote reference 5.

can be called the skewness and kurtosis reduction factors.

It is seen from Fig. 3 and from the limiting values in Table I that the best results are obtained for small values of  $\gamma$ , that is, where the integrator approaches the ideal linear one. A value of  $\gamma = 0$ , while attractive from

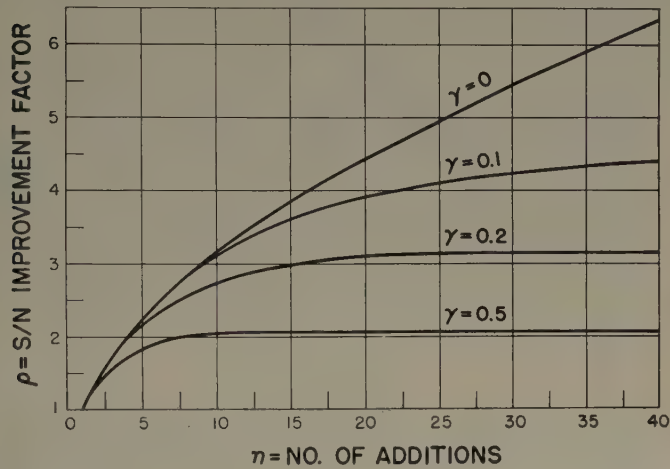


Fig. 3—Signal-to-noise improvement factor,  $\rho$  versus number of additions  $n$  for the nonideal integrator case.

the viewpoint of improvement factor, is manifestly impossible since, in the case of the resonant circuit and delay-line integrators, instability occurs, and in the case of the  $RC$  integrator and storage tube the output would

TABLE I  
LIMITING VALUES

	$\gamma \rightarrow 0$	$\gamma \rightarrow \infty$	$n \rightarrow \infty; \gamma \leq 1/4$
$\rho_1$	$(n)^{1/2}$	1	$(2/\gamma)^{1/2}$
$\rho_3$	$(n)^{-1/2}$	1	$0.943(\gamma)^{1/2}$
$\rho_4$	$1/n$	1	$\gamma$

become vanishingly small, with a  $\gamma = 0$  in the latter case normally signifying a zero current (see (20) et seq.). However, for a given number of signal additions, finite values of  $\gamma$  exist which give results close to those obtained from direct linear addition. For example, from Fig. 3, for  $n = 10$ ,  $\gamma = 0.1$  gives  $\rho = 3.04$  as compared to the 3.16 value ideally obtainable.

To find out just how nonideal an integrator can be and still be worthwhile using, it is helpful to define some quantities equivalent to efficiency factors for comparing the actual to the ideal integrator. For a value of  $\gamma \leq 1/4$  which is the only condition of any great interest to us we find

$$\begin{aligned}\eta_1(n\gamma) &= \rho_1(n, \gamma) / \rho_1(n, 0) = [(\tanh n\gamma/2) / (n\gamma/2)]^{1/2} \\ \eta_3(n\gamma) &= \rho_3(n, \gamma) / \rho_3(n, 0) \cong (n\gamma / \tanh n\gamma)^{1/2} \\ \eta_4(n\gamma) &= \rho_4(n, \gamma) / \rho_4(n, 0) = n\gamma / \tanh n\gamma.\end{aligned}\quad (28)$$

These are rather useful quantities in determining what the permissible value for  $\gamma$  is for a given number of additions. It can be seen that a reciprocal relationship between  $n$  and  $\gamma$ , (i.e.,  $n\gamma = 1$ ) leads to an integrator giving a signal-to-noise improvement ratio 96 per cent of the

maximum attainable and yielding an output probability density with a 14-per cent greater coefficient of skewness and 31-per cent greater excess. Since  $\zeta_3$  and  $\zeta_4$  are, in general, small anyway, these values are quite acceptable for most cases and we can say that, if  $n\gamma = 1$ , we have an integrator which is very nearly an ideal one.

While the foregoing results apply to a variety of practical integrators, they are particularly useful in arriving at the proper operating point in storage-tube integrators. In this instance,  $\gamma$  can be directly measured as the natural logarithmic decrement of the successive-reading signal amplitudes (see Fig. 4).



Fig. 4—Picture of storage-tube "read-out" sequence showing exponential decay.

#### IV. STORAGE-TUBE EXPERIMENTAL RESULTS

A block diagram showing the arrangement of the experimental equipment appears as Fig. 5, in which the major circuitry components are displayed.

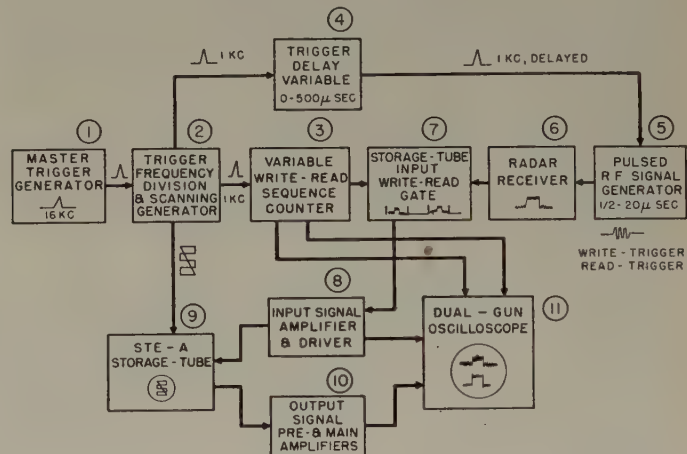


Fig. 5—Block diagram of storage-tube-integrator experimental equipment arrangement.

A master trigger at approximately 16 kc is derived from a sine wave generator, 1.<sup>14</sup> This trigger is used to set the line frequency of the storage tube deflection scan, 2, each line, of which there are sixteen, being 62.5 microseconds in length. The master triggers are also counted down by sixteen to 1 kc, the scanning frame rate. These 1-kc triggers are used to time the variable write-read sequence counter, 3, which provides the proper number of triggers for the number of signals desired to be "written in" to the storage tube, and the proper delay "read-

<sup>14</sup> The numbers refer to the corresponding block of the diagram in Fig. 5.

out" time, after the write interval, during which the signals are taken from the storage tube and the dielectric returned to equilibrium.

The 1-kc trigger is also led through a variable delay, 4, and used to trigger a pulsed 10-centimeter signal generator, 5, whose output pulse may be varied in duration from 0.5 to 20 microseconds. This pulse signal is fed into a radar receiver type APG-5, 6, at which point, by varying the signal generator output power and the radar receiver gain, a  $0 \leq S/N \leq 100$  can be obtained with the input circuits of the receiver furnishing the noise power. The repetitive signals and the accompanying noise are then gated, 7, into the variable bandwidth input amplifier to the STE-A storage tube's back plate an arbitrary number of times and then prevented from reaching the back plate by gating them off during the "read" interval. This signal-plus-noise is also displayed on one channel of a dual-gun oscilloscope, 11; this channel receives a "write" trigger from the write-read sequence counter, 3.

After the desired number of signal-plus-noise sweeps have been placed into storage, the composite signal is gated off the back plate and "read" signals appear at the collector output of the storage tube where they are amplified, 10, and displayed on the second channel of the dual gun oscilloscope; this channel is triggered by the "read" trigger from 3. On the dual-gun oscilloscope, then, there appears on separate sweeps both the composite picture of the signal plus noise before being placed into storage, and the storage tube output; the improvement in signal-to-noise ratio may, therefore, be viewed directly. It should be pointed out that with this experimental procedure erroneous conclusions can be reached regarding the attained  $S/N$  improvement, if the  $n$  reading operations do not give essentially complete erasure, and a value of  $n$  must be chosen such as to erase the dielectric during the read sequence.

With this experimental procedure, improvements in  $S/N$  (for original  $S/N \geq 1$ ) have been investigated as a function of  $n$ , the number of signal-plus-noise additions, and  $\gamma$ , the exponential weighting factor. The range of  $\gamma$ 's explored was between 0.1 and 2, and  $2 \leq n \leq 16$ . Fig. 6 illustrates the theoretical improvement  $\rho$  in  $S/N$  for linear integration ( $\rho = [n]^{1/2}$ ), the theoretical improvement for various values of  $\gamma$ , and the actual improvements measured after storage-tube integration with these values. Fig. 7 is a picture of the improvement for  $n=9$ . Measurements made within the practical limitations of tube operability have borne out the analytical work very well, the improvements predicted have been achieved, and the noise distribution shifts expected have been observed.

The improvement of  $S/N$  in the STE-A storage tube departs in practice from the theory developed in earlier sections because of any or all of the following characteristics of tube operation:

(a) The change in  $\delta$ , the secondary-emission ratio of the dielectric, as a function of excursion from the equi-

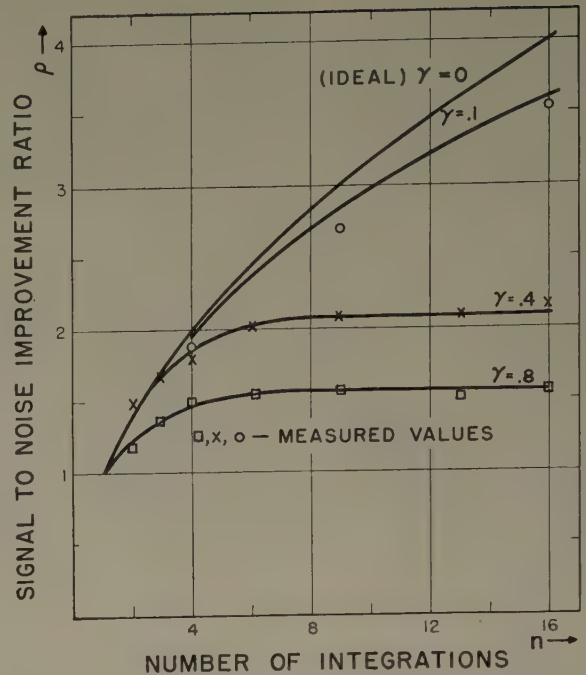


Fig. 6—Measured signal-to-noise improvement  $\rho$  as a function of the number of additions  $n$ .

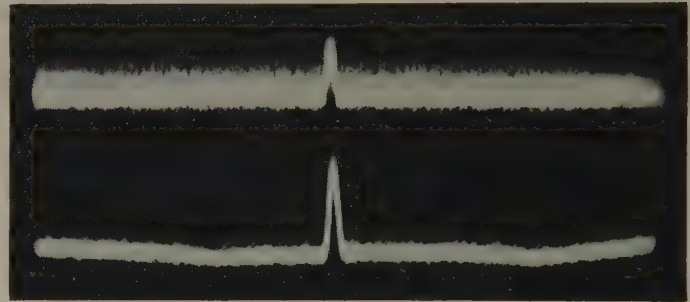


Fig. 7—Pictures of actual signal-to-noise improvement for a 1-microsecond pulse,  $\gamma=0.1$  and  $n=9$  additions.

librium potential is approximately linear over only a relatively small range of voltage changes applied to the back plate. As a consequence of using only small input voltages to the back plate in order to stay within this linear region, the dielectric cannot assume full charge and the output signal is small; for these small output signals, noise in the first stage of the wide-band output amplifier will diminish the expected improvement in  $S/N$ .

(b) For maximum improvement a vanishingly small value of  $\gamma$  is desirable. From the definition of  $\gamma$  ((20) et seq.), however, it is seen that  $K$  and  $C$  cannot be varied without making changes internal to the tube, and since  $V$  is fixed once the signal repetition interval has been defined, the only parameter capable of being varied is  $I_b$ . Inasmuch as  $\gamma \propto I_b$ , a small  $I_b$  is called for; this results in the same condition as (a) above, however, in that as  $I_b$  is reduced, the output signal becomes smaller and output amplifier noise becomes a limiting factor to  $S/N$  improvement. Additional circuitry (a positive gat-

ing voltage applied to the grid) might be employed to increase the beam current during the read interval, however, thus reducing this limitation somewhat.

(c) Even under static equilibrium conditions, when there is no signal being placed into or taken from storage, a periodic "disturbance" appears at the output as the electron beam is swept across the dielectric. (See, for instance, Fig. 4.) This disturbance has line and frame frequency components and arises, for the most part, from the change in collection efficiency as a function of angular departure of the primary beam current from the normal to the dielectric. With the limitations on input signal voltage mentioned in (a) above, this disturbance sets an upper limit of about 10 to 15 (depending upon the particular tube) on the signal-to-disturbance ratio viewed at the output of the tube.

## V. CONCLUSIONS

The theory of the ideal (linear) and nonideal integrator has been developed, relating improvement in signal-to-noise ratio to such parameters as original signal-to-noise ratio, detector function, number of additions and the degree of integrator nonlinearity. The improvements expected have been verified experimentally by using a storage tube as the integrating device. At least where the number of integrations is small ( $n \leq$  about 25) and the output  $S/N$  does not need to be higher than about 10 to 15, the STE-A barrier-grid storage tube is shown to possess practical utility as a repetitive signal integrator.

## ACKNOWLEDGEMENT

The author wish to acknowledge the helpful advice and criticism received from A. S. Jensen of the RCA Princeton Laboratories, E. W. Samson, the chief of the Communications and Relay Laboratory in which the work was carried on, and many members of the Air Force Cambridge Research Laboratories.

## GLOSSARY

(Terms not constants or not defined explicitly in the text appear here.)

$C$  = capacitance

${}_1F_1$  = confluent hypergeometric function

$H_n$  =  $n$ th Hermite polynomial

$I_0$  = modified Bessel function of the first kind and zero order

$I_b$  = beam current

$I_N$  = noise current

$L$  = inductance

$n$  = an integer; also, number of integrated signals

$P$  = sine wave amplitude

$R$  = envelope of sine wave plus noise; also, resistance

$\bar{R}^n$  =  $n$ th moment of a distribution  $R$

$S$  = generalized frequency

$V$  = beam-scanning velocity

$X$  = a general variable

$\Gamma$  = the gamma function

$\mu_k$  = the  $k$ th moment of a distribution

$\omega = 2\pi f$

$\psi_0$  = noise power.

# Distortion—Band-Pass Considerations in Angular Modulation\*

ALBERT A. GERLACH†, ASSOCIATE, IRE

**Summary**—The distortion introduced into an angular modulated signal transmitted through a network or medium is analyzed using the sideband approach. An exact open-form solution of the output signal is derived; however, the usefulness of this equation is limited by the ability to perform the indicated integrations. It is shown that exact solutions may readily be found for transfer functions which are linear exponential functions of frequency, and since over a given frequency interval, any transfer function may be expanded in an orthogonal set of linear exponential functions of frequency, a solution is always obtainable. A few examples of possible transfer functions are treated, illustrating the distortion tendencies for both the

amplitude and phase characteristics. A rule-of-thumb formula is derived to determine the maximum undistorted modulation frequency which may be transmitted through a network of a given band-pass characteristic. It is concluded that linear phase characteristics are a more important design objective than flat amplitude characteristics in minimizing distortion.

## INTRODUCTION

WITH THE INCREASING USE of frequency modulation for instrumentation purposes, more concern is being focused on the nature and quantity of distortion introduced in the intelligence during transmission. A topic of much study has been to determine accurately the output intelligence of an angular

\* Decimal classification: R148.2. Original manuscript received by the Institute, August 15, 1949; revised manuscript received, March 31, 1950. Presented, 1950 IRE National Convention, New York, N. Y., March 7, 1950.

† Armour Research Foundation of Illinois Institute of Technology, Technology Center, Chicago, Ill.

modulated signal which is transmitted through a medium of a given sinusoidal-frequency transfer characteristic. (A transfer function is that expression which relates an output quantity to an input quantity as a function of the sinusoidal frequencies of the two quantities.) Unfortunately, this problem is a very complex one and it becomes exceedingly difficult to solve the problem for the general transfer function. As a matter of fact, no closed mathematical expression has been derived which exactly relates the output signal to the input signal for the general transfer function. What has been done is to express the output signal as a function of the input signal and the general transfer function in the form of a differential equation, a definite integral, or an infinite series; or to approximate the output signal in the form of a finite series of terms. The versatility of this latter method is, of course, restricted by the approximations made.

### HISTORICAL REVIEW

There have been two general methods of attacking the problem although there are many variations of these methods. These methods are the instantaneous signal approach, keeping the signals in their exact form as a function of time and using the transfer function as a differential operator to derive the output signal, and the spectrum analysis approach where the input signal is transformed into a sinusoidal frequency spectrum and the output signal is then the sum of the products of the individual components and the transfer function at that frequency.

Carson and Fry<sup>1</sup> have expressed the output signal as an infinite series, the first term of which has been called the quasi-steady-state result and is valid only within a restricted range of the deviation, modulation frequency and bandwidth of the network. Jaffe,<sup>2</sup> Van der Pol,<sup>3</sup> Gladwin,<sup>4</sup> and others have made use of, or arrived at the same results by other methods. On the other hand, Frantz<sup>5</sup> and Gold<sup>6</sup> have exploited the sideband or spectrum analysis method to arrive at formulas for a general result, limited again by a more well-defined approximation than the quasi-steady-state method.

All of the methods of analysis to date have followed the more or less logical procedure of requiring the analysis to be a slave to the transfer function which is carried along in as general a condition as is possible. In the analysis to follow, the reverse procedure has been em-

ployed; that is, a search has been made for those transfer characteristics which would result in a manageable and exact solution of the output quantity. Ordinarily this approach to the problem would lead to pure abstract relations which would have little bearing on a practical solution; however, as shall be shown, the results obtained are quite useful in predicting the distortion effects on the signal intelligence for most practical transfer characteristics.

### SOME BASIC EQUATIONS

In the following paragraphs will be derived some basic equations for the solution of an output quantity when an input quantity or signal is applied to a medium or network of known transfer function.

The general angular modulated signal may be expressed as

$$e(t) = E \sin [\omega_0 t + \theta(t)] \quad (1)$$

where  $\omega_0$  is the carrier frequency about which the signal intelligence function varies. (In a phase-modulation system the intelligence will be directly proportional to  $\theta(t)$ , and in a frequency-modulation system the intelligence will be proportional to the time derivative of  $\theta(t)$ .)

Keeping in mind that it is the imaginary part that is of interest, (1) may be written

$$e(t) = E e^{i[\omega_0 t + \theta(t)]} = E e^{i\omega_0 t} e^{i\theta(t)}. \quad (2)$$

If the function  $\theta(t)$  is periodic of period  $T$  it may be expanded in a Fourier series in terms of the orthogonal set of functions  $e^{in\rho t}$  as follows:<sup>7</sup>

$$e^{i\theta(t)} = \sum_{-\infty}^{\infty} C_n e^{in\rho t} \quad (3)$$

where

$$\rho = 2\pi/T$$

and

$$C_n = \frac{1}{2\pi} \int_{-\pi}^{\pi} e^{i\theta} e^{-in\rho t} d\rho t = \frac{1}{2\pi} \int_{-\pi}^{\pi} e^{i(\theta - n\rho t)} d\rho t. \quad (4)$$

(If  $\theta$  is a sinusoidal function of frequency, (4) may be recognized as the Bessel integral form of the Bessel function of the first kind of order  $n$ .) Equation (2) may then be expressed as

<sup>1</sup> J. R. Carson and T. C. Fry, "Variable electric circuit theory," *Bell Sys. Tech. Jour.* vol. 16, p. 513; October, 1937.

<sup>2</sup> D. L. Jaffe, "A theoretical and experimental investigation of tuned circuit distortion in frequency-modulation systems," *Proc. I.R.E.*, vol. 33, pp. 318-334; May, 1945.

<sup>3</sup> B. Van der Pol, "Fundamental principle of frequency modulation," *Jour. IEE*, Part III, vol. 93, pp. 153-158; May, 1946.

<sup>4</sup> A. S. Gladwin, "The distortion of frequency-modulated waves by transmission networks," *Proc. I.R.E.*, vol. 35, pp. 1436-1445; December, 1947.

<sup>5</sup> W. J. Frantz, "The transmission of a frequency-modulated wave through a network," *Proc. I.R.E.*, vol. 34, pp. 114P-125P; March, 1946.

<sup>6</sup> B. Gold, "The solution of steady-state problems in FM," *Proc. I.R.E.*, vol. 37, pp. 1264-1269; November, 1949.

<sup>7</sup> It is assumed here that the function concerned possesses the necessary properties for the Fourier series or Fourier integral, as the case may be, to converge to the mean value of the function over the necessary interval. This does not impose too stringent a restriction on the function since sufficient conditions for the Fourier series representation are that the function be periodic, be sectionally continuous over the fundamental period, and possess right- and left-hand derivatives at all points of the independent variable. Sufficient conditions for the Fourier integral representation are that the function be sectionally continuous in every finite interval, possess right- and left-hand derivatives at all points of the independent variable, and that over the limits minus infinity to plus infinity the integral of the function with respect to its independent variable converges.

$$e(t) = E\epsilon^{i\omega_0 t} \sum_{-\infty}^{\infty} C_n \epsilon^{in\rho t}. \quad (5)$$

If now this quantity is impressed on a medium or network with transfer function  $Z(\omega)$  (see Fig. 1), the output response may be written as

$$e_0(t) = E\epsilon^{i\omega_0 t} \sum_{-\infty}^{\infty} C_n \epsilon^{in\rho t} Z(\omega_0 + n\rho). \quad (6)$$

The procedure up to this point has been to express a complex function of time into an infinite series of sinusoidal components. Each component then is multiplied by the appropriate transfer impedance at that particular frequency and the resultant sum of all of these products is then the output response as indicated by (6) above. Now it is possible by the Fourier transform to express the transfer function  $Z(\omega)$  in a definite integral form as<sup>7</sup>

$$Z(\omega) = \int_{-\infty}^{\infty} F(\tau) \epsilon^{-i\omega\tau} d\tau \quad (7)$$

where

$$F(\tau) = \frac{1}{2\pi} \int_{-\infty}^{\infty} Z(\omega) \epsilon^{i\omega\tau} d\omega. \quad (8)$$

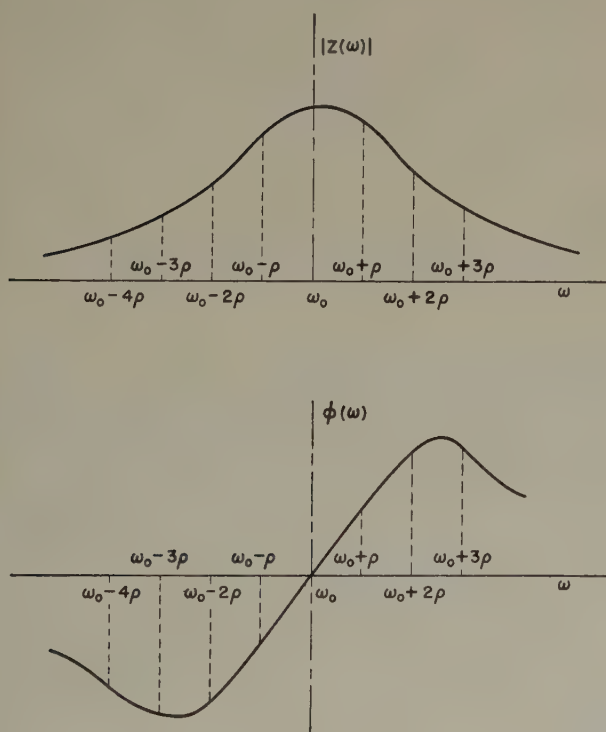


Fig. 1

Equation (6) may then be written

$$\begin{aligned} e_0(t) &= E\epsilon^{i\omega_0 t} \int_{-\infty}^{\infty} F(\tau) \epsilon^{-i\omega_0 \tau} \sum_{-\infty}^{\infty} C_n \epsilon^{in\rho(t-\tau)} d\tau \\ &= \int_{-\infty}^{\infty} F(\tau) e(t-\tau) d\tau \\ &= \frac{1}{2\pi} \int_{-\infty}^{\infty} e(t-\tau) \int_{-\infty}^{\infty} Z(\omega) \epsilon^{i\omega\tau} d\omega d\tau, \end{aligned} \quad (9)$$

when  $\theta(t)$  is a sinusoidal function; that is, let

$$\theta(t) = m \sin \rho t. \quad (10)$$

Equation (9) may be expressed as

$$e_0(t) = E\epsilon^{i\omega_0 t} Y(\omega_0, t) \quad (11)$$

where

$$Y(\omega_0, t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \epsilon^{im \sin \rho(t-\tau)} \int_{-\infty}^{\infty} Z(\omega) \epsilon^{i(\omega-\omega_0)\tau} d\omega d\tau. \quad (12)$$

It is the argument of the above expression which contains the output intelligence, however, the magnitude of  $Y(\omega_0, t)$  is also of some interest since it will indicate the degree of limiting necessary to eliminate the undesirable amplitude variations.

It is of interest to note that if the periodicity of  $\theta(t)$  is not assumed, an equally general solution for  $e_0$  may be obtained by making use of the Fourier transforms. This expression may readily be shown to be

$$e_0(t) = \frac{E}{2\pi} \int_{-\infty}^{\infty} Z(\omega) \int_{-\infty}^{\infty} \epsilon^{i[\omega_0 \tau + \theta(\tau)]} \epsilon^{i\omega(t-\tau)} d\tau d\omega. \quad (13)$$

The procedure will now be to determine the nature of the transfer functions which will allow the output quantity to be expressed in a closed mathematical form and to determine what conclusions may be drawn from these functions about the general case.

Rather than investigate possible solutions of the integral in (9) directly, it will be simpler and more instructive to work from (6). It may be noticed from this equation that transfer functions which are linear exponential functions of frequency may be combined with the Fourier series so that this series is directly transformable into its original closed form with the time variable shifted by a constant factor dependent on the nature of the exponent of the transfer function. This property was utilized in (9) in the derivation of the general output quantity. Let

$$Z(\omega) = \epsilon^{\alpha + \beta\omega}. \quad (14)$$

From (6) then the output response may be written

$$\begin{aligned} e_0(t) &= E\epsilon^{\alpha + i\omega_0(t-i\beta)} \sum_{-\infty}^{\infty} C_n \epsilon^{in\rho(t-i\beta)} \\ &= \epsilon^{\alpha} e(t - i\beta). \end{aligned} \quad (15)$$

It will be of considerable interest to note that the entire result may be generalized for any function<sup>7</sup> of time  $e(t)$  regardless of whether or not the function is periodic or of the form indicated by (1). The procedure is as follows:

$$e_0(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \epsilon^{\alpha + \beta\omega} G(\omega) \epsilon^{i\omega t} d\omega \quad (16)$$

where

$$G(\omega) = \int_{-\infty}^{\infty} e(t) \epsilon^{-i\omega t} dt. \quad (17)$$

Therefore

$$e_0(t) = \frac{\epsilon^\alpha}{2\pi} \int_{-\infty}^{\infty} G(\omega) \epsilon^{i\omega(t-i\beta)} d\omega = \epsilon^\alpha e(t - i\beta). \quad (18)$$

Now, over the range of frequencies where the spectral energy of  $e(t)$  is appreciable (see (17)), any transfer function<sup>7</sup> may be represented by a Fourier series. The series will generally not represent the transfer function elsewhere; however, this is of little consequence since no appreciable signal energy exists elsewhere. Let  $B$  be the bandwidth in cycles containing the appreciable spectral energy and  $\omega_0$  be the angular frequency of the center of the spectral range, then over the range  $\omega - B/2 \leq \omega \leq \omega + B/2$ ,  $Z(\omega)$  may be represented by<sup>7</sup>

$$Z(\omega) \sim \sum_{-\infty}^{\infty} a_n \epsilon^{in(\omega/B)} \quad (19)$$

where

$$a_n = \frac{1}{2\pi B} \int_{\omega_0 - B/2}^{\omega_0 + B/2} Z(\omega) \epsilon^{-in(\omega/B)} d\omega. \quad (20)$$

Equation (19) has the form of (14) if one recognizes the fact that

$$\alpha_n = \log a_n \quad \text{and} \quad \beta_n = i \frac{n}{B}. \quad (21)$$

The output signal may then be written directly from (15) or (18) as

$$e_0(t) = \sum_{-\infty}^{\infty} a_n e\left(t + \frac{n}{B}\right). \quad (22)$$

By extension of the above analysis using the Fourier transforms it may be shown that the following relation between the input and output functions is valid (under rather moderate restrictions):

$$e_0(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} e(t - \tau) \int_{-\infty}^{\infty} Z(\omega) \epsilon^{i\omega\tau} d\omega d\tau. \quad (23)$$

Equation (23) should prove to be a useful formula in solving  $e_0(t)$  for any desired transfer function  $Z(\omega)$ . It is quite interesting in that it provides a direct solution of the output voltage as a function of time in terms of the input voltage as a function of time and the transfer function as a function of frequency. Equation (23) may still be useful if the first integral does not converge provided one considers  $Z(\omega)$  to extend over finite limits sufficient to include the appreciable spectral energy of  $e(t)$  and to be zero for all other values of the variable  $\omega$ .

#### A FEW EXAMPLES

As a first example of the use of (18) and (22) consider the case of flat-amplitude transfer characteristics and linear-phase transfer characteristics. In this case,  $\alpha$  may be complex and  $\beta$  will be pure imaginary and, therefore,

the signal intelligence  $\theta(t)$  will just experience a time delay through the transmission medium.

Consider next that the phase transfer characteristics are linear, and that over the portion of the frequency spectrum where the spectral components of the angular modulated signal are appreciable, the amplitude obeys the following function:

$$|Z(\omega)| = \cosh \lambda(\omega - \omega_0) = \frac{1}{2} [\epsilon^{\lambda(\omega - \omega_0)} + \epsilon^{-\lambda(\omega - \omega_0)}]. \quad (24)$$

(See Fig. 2.) From (2) and (18) then

$$e_0(t) = \frac{1}{2} E \epsilon^{i\omega_0 t} [\epsilon^{i\theta(t-i\lambda)} + \epsilon^{i\theta(t+i\lambda)}]. \quad (25)$$

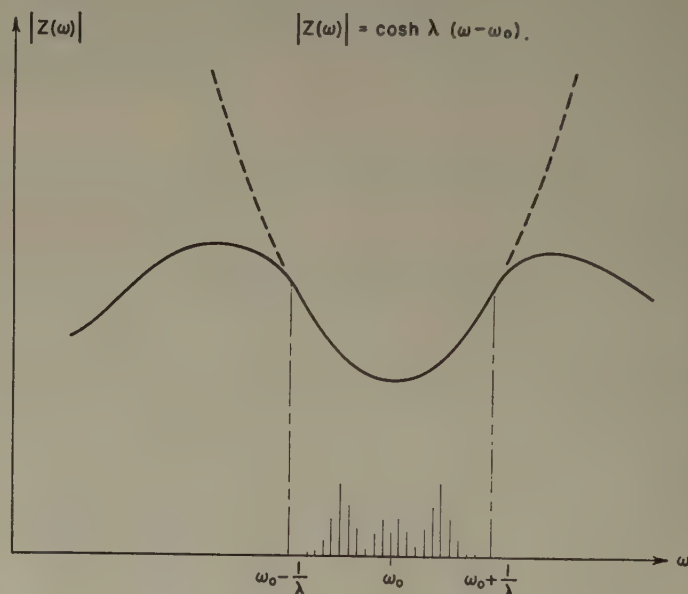


Fig. 2

If  $\theta(t)$  is a sinusoidal function (see (10)), (25) becomes

$$e_0(t) = E \cosh [(m \sinh \rho \lambda) \cos \rho t] \epsilon^{i[\omega_0 t + (m \cosh \rho \lambda) \sin \rho t]}. \quad (26)$$

Equation (26) illustrates the rather interesting result that the signal intelligence suffers no nonlinear distortion and, additionally, that the higher signal frequency components are exalted by an amount  $\cosh \rho \lambda$ . It therefore becomes conceivable that this property may be used advantageously in a frequency-modulation system to enhance the higher frequency components. The chief limitation of this property is that the greater the signal frequency emphasis the greater also will the magnitude variation of  $e_0(t)$  requiring, therefore, increased limiting of the output signal.

As a third example, consider the case where the phase characteristics are again either zero or essentially linear over the pertinent portion of the frequency spectrum and that the amplitude characteristics obey the function

$$|Z(\omega)| = \cos (\lambda \omega + \Omega) = \frac{1}{2} [\epsilon^{i(\lambda \omega + \Omega)} + \epsilon^{-i(\lambda \omega + \Omega)}]. \quad (27)$$

(See Fig. 3.) From (2) and (18) then

$$e_0(t) = \frac{1}{2} E e^{i\omega_0 t} \left\{ e^{i[(\Omega + \omega_0\lambda) + \theta(t+\lambda)]} + e^{-i[(\Omega + \omega_0\lambda) - \theta(t-\lambda)]} \right\}. \quad (28)$$

If now  $\theta(t)$  is a sinusoidal function (see (10)), (28) becomes

$$e_0(t) = E \cos [(m \sin \rho\lambda) \cos \rho t + (\Omega + \omega_0\lambda)] e^{i[\omega_0 t + (m \cos \rho\lambda) \sin \rho t]}. \quad (29)$$

Equation (29) again illustrates that the intelligence suffers no nonlinear distortion; however, in this case the higher frequency signals are depressed to an amount equal to  $\cos \rho\lambda$  of the original amplitude. This fact is of primary interest in the application of angular modulation techniques to magnetic recording since, to a first approximation, the amplitude playback characteristics of a tape or wire recorder is a sine wave.

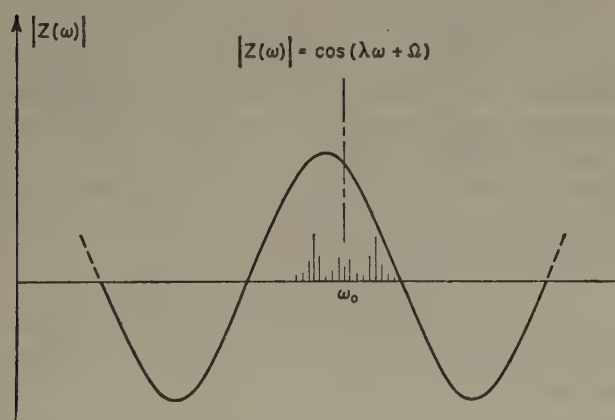


Fig. 3

Consider now that  $B$  is the band pass in cycles of the 3-db down points of the cosine transfer function and let  $\delta$  be the maximum amount of signal depression that can be tolerated; then it may readily be shown that the relationship between  $B$  and the maximum signal intelligence frequency that may be transmitted is

$$f_s = \frac{2B}{\pi} \cos^{-1} (1 - \delta). \quad (30)$$

For small values of  $\delta$  this may be approximated as

$$f_s = 0.9B\sqrt{\delta}. \quad (31)$$

The sinusoidal-amplitude transfer characteristic therefore imposes quite a fundamental limitation on the maximum modulation frequency which may be transmitted, based solely on the shape of the transfer characteristic and completely independent of the deviation frequency. For example if  $\delta$  is two per cent, the required band-pass  $B$  should be about eight times the maximum modulation frequency.

## CONCLUSIONS

The output of an angular modulated signal transmitted through a network or medium (under certain restrictions<sup>7</sup>) may be expressed as a double definite integral (see (9), (11), (13), and (23)). The usefulness of these open-form equations is limited by the ability to perform the indicated integrations. It was found, however, that transfer functions which are linear exponential functions of frequency readily lend themselves to a closed mathematical solution of the output signal. Since any transfer function<sup>7</sup> may be expanded in a Fourier series of linear exponential functions of frequency over a desired range of frequencies, the output signal may always be expressed in a finite number of terms, the number of terms being the number of terms of the Fourier expansion necessary to approximate the transfer function over the range of frequencies where the spectral components of the input signal are appreciable. Equations (18), (22), and (23) are more inclusive formulas in that they express an output function of time in terms of a general input function of time and the transfer function relating the output and input as a function of frequency.

A few particular cases were investigated in order to study the effects of the amplitude and phase characteristics of the transfer function on the output intelligence. It was found that for linear phase characteristics, the output intelligence will suffer no distortion but will be shifted along the time axis by an amount depending on the slope of the phase characteristic. If the amplitude characteristics are a hyperbolic cosine function of frequency the output intelligence will suffer no nonlinear distortion, but the amplitude of the signal intelligence will be an increasing function of frequency. On the other hand, if the amplitude transfer characteristics are a sinusoidal function of frequency there will again be no nonlinear distortion in the output intelligence; however, in this case, the amplitude of the intelligence will be a decreasing function of frequency. This amplitude characteristic imposes a fundamental limitation on the maximum frequency response of the intelligence signals as is indicated by (30). By extension of the analysis it may be shown that, generally speaking, amplitude transfer characteristics which rise on either side of the carrier frequency enhance the higher modulation frequencies, while those which fall on either side of the carrier frequency depress the higher modulation frequencies. In this latter case, (30) may be used as a rule-of-thumb formula to determine the maximum modulation frequency which may be transmitted without appreciable distortion.

This analysis again supports the contention and conclusions of other writers on the subject that linear-phase transfer characteristics are a more important design objective than are perfectly flat-amplitude transfer characteristics.

# Pulse Transients in Exponential Transmission Lines\*

EDWARD R. SCHATZ†, ASSOCIATE, IRE, AND EVERARD M. WILLIAMS‡, SENIOR MEMBER, IRE

**Summary**—The pulse response of exponential transmission lines is analyzed and it is suggested that such lines may be used advantageously as pulse transformers with short, rapidly rising pulses, particularly where large amounts of power are involved. A subsequent paper will discuss design problems and experimental results.

THE EXPONENTIAL transmission line shows promise as a pulse transformer for use with very short, rapidly rising pulses, particularly where large amounts of power are involved. The design and effective use of exponential transmission-line sections as pulse transformers require a fairly precise knowledge of the general transient characteristics of such lines. It is the purpose of this paper to discuss some of these general characteristics. A subsequent paper will treat design problems and describe some typical exponential-line pulse transformers.

Requirements exist for pulse-transforming devices capable of handling large amounts of power in pulse durations of the order of hundredths of microseconds; examples of such requirements are to be found, for instance, in the "electrostatic" ion deflectors for large nuclear particle accelerators. Such requirements are not readily satisfied by conventional two-winding iron-core transformers and, in a search for an alternative wide-band impedance-transforming device, the authors were led to consider the exponential transmission line. The steady-state characteristics of exponential transmission lines have been widely discussed in the literature<sup>1-3</sup> and it is known that their behavior is equivalent to that of combinations of high-pass filters and ideal impedance transformers. The high-pass characteristic is an ideal one for the transformation of steep edges of pulses but necessarily distorts their flat-topped portions. It seemed probable, however, that the parameters of an exponential transmission line could be so chosen as to reduce this distortion to any desired extent. The analytical study described in this paper was, therefore, undertaken to investigate the transient properties of the exponential transmission line.

An exponential line is defined as a transmission line in which spacing between conductors is tapered so that the

inductance and capacitance per unit length  $l_x$  and  $c_x$  vary exponentially with the distance along the line from the sending end according to the relations

$$l_x = L_0 e^{\gamma x}$$

$$c_x = C_0 e^{-\gamma x}.$$

The flare coefficient  $\gamma$  may be either positive or negative; the analysis of this paper, however, considers positive values of  $\gamma$  only. The differential equations for  $V_x$  and  $i_x$  are, by an analysis similar to that for uniform lines,

$$\frac{\partial^2 V_x}{\partial x^2} - \gamma \frac{\partial V_x}{\partial x} = L_0 C_0 \frac{\partial^2 V_x}{\partial t^2} \quad (1)$$

$$\frac{\partial^2 i_x}{\partial x^2} + \gamma \frac{\partial i_x}{\partial x} = L_0 C_0 \frac{\partial^2 i_x}{\partial t^2} \quad (2)$$

A method of solution using Laplace transform analysis was employed; the general steps in this solution are illustrated briefly for (1).

The Laplace transform  $F(x, s)$  of a function  $f(x, t)$  is defined as

$$F(x, s) = \int_0^\infty e^{-st} f(x, t) dt$$

and the operation of taking the Laplace transform is indicated as

$$F(x, s) = Lf(x, t).$$

The notations  $V(x, s)$  and  $I(x, s)$  will be used for the transforms of  $v(x, t)$  and  $i(x, t)$ . Taking the Laplace transforms of each term in (1) we obtain

$$L \frac{\partial^2 v(x, t)}{\partial x^2} - \gamma L \frac{\partial v(x, t)}{\partial x} = L_0 C_0 L \frac{\partial^2 v(x, t)}{\partial t^2}$$

which for the conditions of this analysis can be shown to become

$$\frac{\partial^2 V(x, s)}{\partial x^2} - \gamma \frac{\partial V(x, s)}{\partial x} = L_0 C_0 s^2 V(x, s).$$

The solution of this is

$$V = e^{\gamma x/2} [A e^{-\sqrt{(\gamma^2/4) + s^2 L_0 C_0} x} + B e^{\sqrt{(\gamma^2/4) + s^2 L_0 C_0} x}] \quad (3)$$

in which  $A$  and  $B$  are functions of  $s$  and independent of  $x$ .

The first and second terms of (3) represent traveling waves in the  $x$  and  $-x$  directions, respectively. The parameter  $\sqrt{1/L_0 C_0}$ , hereafter designated as  $c$ , is the velocity of propagation of electromagnetic waves in the space between conductors. The solution is completed by using boundary conditions to solve for the terms  $A$  and  $B$ . This procedure yields  $V(x, s)$ . The desired voltage  $v(x, t)$  is obtained from the inverse transform of  $V(x, s)$ .

\* Decimal classification: R117.1. Original manuscript received by the Institute, January 3, 1949; revised manuscript received, April 24, 1950. This paper is part of a dissertation submitted by Edward R. Schatz in partial fulfillment of the requirements for the degree of Doctor of Science at Carnegie Institute of Technology. Also presented, in part, under the title "The exponential line pulse transformer," 1950 IRE National Convention, New York, N. Y., March 9, 1950.

This work was supported in part by the joint ONR-AEC program. † Department of Electrical Engineering, Carnegie Institute of Technology, Pittsburgh, Pa.

<sup>1</sup> H. A. Wheeler, "Transmission lines with exponential taper," *Proc. I.R.E.*, vol. 27, pp. 65-71; January, 1939.

<sup>2</sup> C. R. Burrows, "The exponential transmission line," *Bell Sys. Tech. Jour.*, vol. 17, pp. 555-573; October, 1938.

<sup>3</sup> P. J. Selgin, "Electrical Transmission in the Steady State," McGraw-Hill Book Co., New York, N. Y., 1946.

This operation of finding the inverse transform is extremely complicated in the case of a general arbitrary termination; the authors have obtained solutions for the cases summarized in Table I.

The first case is the infinite line. The integrals appearing in this case have been computed for the conditions shown in the curves of Figs. 1 and 2. Fig. 3 is a sketch of the voltage and current distribution on an infinite line, excited by a short rectangular pulse, for four instants of time at which the leading edge of the pulse has reached the positions  $x_1, x_2, x_3$ , and  $x_4$ , respectively.

Sending end current as a function of time is given in the second case. Fig. 4 shows a curve of this function, useful in determining the sending end resistance and in efficiency calculations.

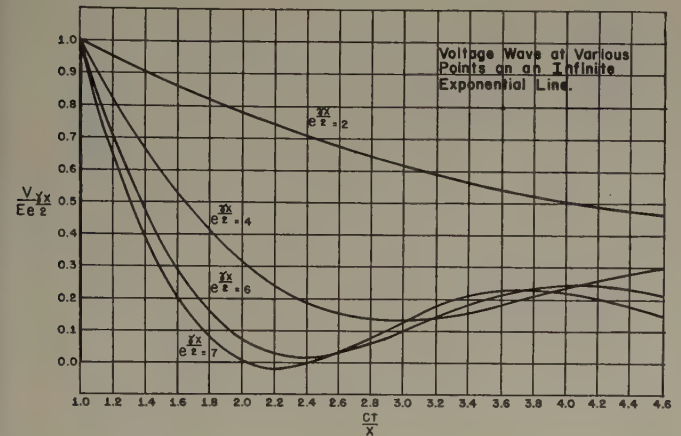


Fig. 1—Voltages at the end of step-up sections of exponential transmission line of length  $x$  (terminated in an indefinite continuation of the same line) as functions of a time parameter  $ct/x$  when the lines are excited with a step input voltage  $E$  at time  $t=0$ . Curves are drawn for sections with voltage ratios of one to two, four, six, and seven, and voltage is expressed as a ratio of actual voltage to the input voltage multiplied by the voltage ratio in each case.

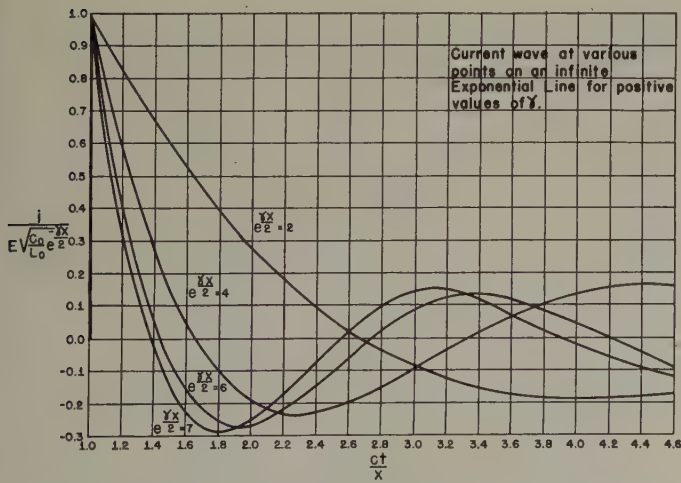


Fig. 2—Currents at the end of voltage step-up sections of exponential transmission line of length  $x$  (terminated in an indefinite continuation of the same line) as functions of the time parameter  $ct/x$ . Curves are drawn for sections with voltage ratios corresponding to those in Fig. 1, and current is expressed as a ratio of actual current to the input current divided by the voltage ratio in each case.

The third case is that of the short-circuited line. The last case is that of the voltage at a receiving end load

$$R_{x_1} = \sqrt{\frac{L_{x_1}}{C_{x_1}}} = \sqrt{\frac{L_0}{C_0}} e^{\gamma x_1}.$$

This resistance is equal to the characteristic impedance at infinite frequency and will be termed the “impedance level” at the output terminals of the length  $x_1$  of line. With this termination there is no reflection from a step-impulse wave front but reflections occur immediately

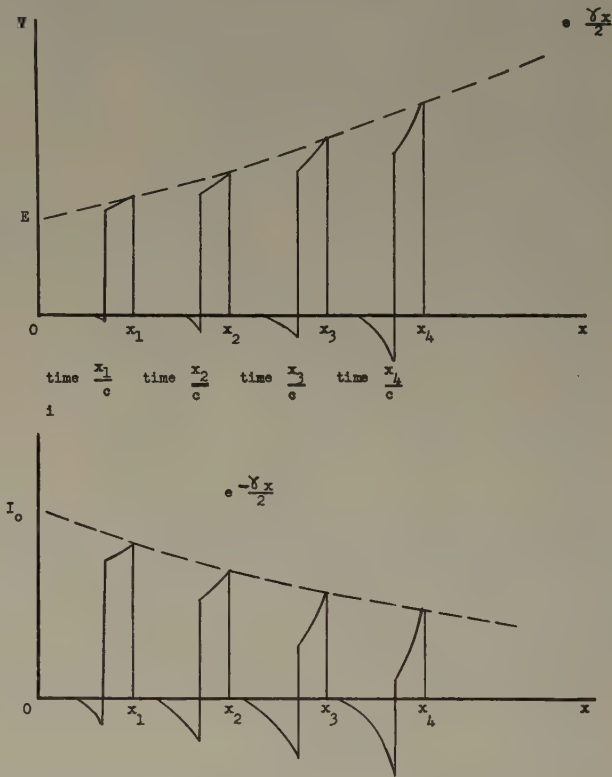


Fig. 3—Sketches of voltage and current distribution on an infinite line, excited by a short rectangular voltage pulse, for several instants of time at which the pulse has reached the positions  $x_1, x_2, x_3$ , and  $x_4$ . The distortion is exaggerated and corresponds to that occurring with very great taper.

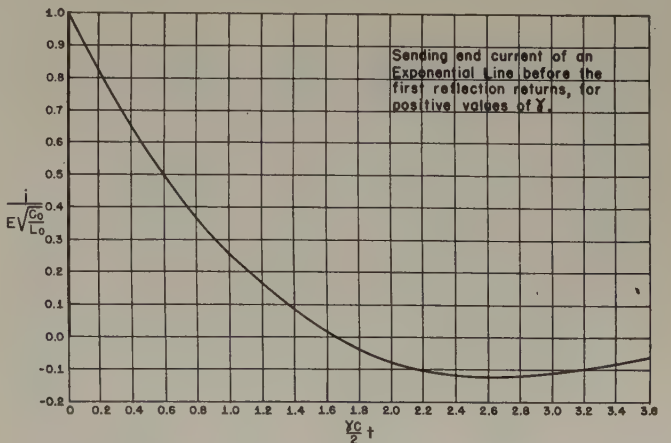


Fig. 4—Current at the sending end of an exponential transmission line when the line is excited at this point by a step voltage function. This current is applicable not only in the case of the line of infinite extent but also to any line section prior to the arrival of reflections from the receiving (load) end.

TABLE I

Voltage  $v$  and current  $i$  as functions of time  $t$  and distance  $x$  along exponential transmission lines excited with a step function of voltage  $E$  at  $x=0$  and  $t=0$ . Solutions are for positive flare coefficient  $\gamma$  only.  $J_0, J_1$  are Bessel functions of the first kind and  $H_0, H_1$  are Struve functions. The notations  $f_1(\tau)$  and  $f_1(t)$  are used for  $\sqrt{\tau^2 - x^2/c^2}$  and  $\sqrt{t^2 - x^2/c^2}$ , and  $N(t-x/c)$  is a function with the value 0 for  $t < x/c$  and unity for  $t > x/c$ .

Case	Voltage transform $V$ or current transform $I$	Solution for voltage or current
Infinite line Voltage and Current at any point $x$	$V(x, s) = \frac{E}{s} e^{-\gamma x/2} \left[ \epsilon^{-\sqrt{(s^2/c^2) + (\gamma^2/4)x}} \right]$ $I(x, s) = \frac{-E}{L_0} e^{-\gamma x/2} \left[ \frac{\gamma}{2s^2} - \sqrt{\frac{1}{s^2/c^2} + \frac{\gamma^2}{4s^4}} \right] \epsilon^{-\sqrt{(s^2/c^2) + (\gamma^2/4)x}}$	$V(x, t) = E e^{\gamma x/2} \left[ 1 - \frac{\gamma x}{2} \int_{x/c}^t \frac{J_1 \left[ \frac{\gamma c}{2} f_1(\tau) \right] d\tau}{f(\tau)} \right] N \left[ t - \frac{x}{c} \right]$ $i(x, t) = \frac{E}{\sqrt{L_0}} e^{-\gamma x/2} \left[ \frac{\gamma^2 c x}{4} \int_{x/c}^t \frac{J_1 \left[ \frac{\gamma c}{2} f_1(\tau) \right] d\tau}{f(\tau)} - \frac{\gamma c}{2} \left( t - \frac{x}{c} \right) + J_0 \left[ \frac{\gamma c}{2} f_1(t) \right] \right. \\ \left. + \frac{\gamma^2 c^2 t}{4} \int_{x/c}^t J_0 \left[ \frac{\gamma c}{2} f_1(\tau) \right] d\tau - \frac{\gamma c}{2} f_1(t) J_1 \left[ \frac{\gamma c}{2} f_1(t) \right] \right] N \left( t - \frac{x}{c} \right)$
Sending end current (for any case) before first re- flection returns	$I(0, s) = \frac{-E}{L_0} \left[ \frac{\gamma}{2s^2} - \sqrt{\frac{1}{s^2/c^2} + \frac{\gamma^2}{4s^4}} \right]$	$i(0, t) = \frac{E}{\sqrt{L_0}} \left[ \left( 1 + \frac{\gamma^2 c^2 t^2}{4} \right) J_0 \left( \frac{\gamma c t}{2} \right) + \pi \left( \frac{\gamma^2 c^2 t^2}{8} \right) \left[ J_1 \left( \frac{\gamma c t}{2} \right) H_0 \left( \frac{\gamma c t}{2} \right) \right. \right. \\ \left. \left. - J_0 \left( \frac{\gamma c t}{2} \right) H_1 \left( \frac{\gamma c t}{2} \right) \right] - \frac{\gamma c t}{2} J_1 \left( \frac{\gamma c t}{2} \right) \right] N \left( t - \frac{x}{c} \right)$
Open-circuited line of length $x_1$ (Receiving end voltage before reflections from sending end)	$V(x_1, s) = E e^{\gamma x_1/2} \left[ \frac{2\epsilon^{-\sqrt{(s^2/c^2) + (\gamma^2/4)x_1}}}{s} \right] \left[ \frac{\sqrt{\frac{s^2}{c^2} + \frac{\gamma^2}{4}}}{\frac{\gamma}{2} + \sqrt{\frac{s^2}{c^2} + \frac{\gamma^2}{4}}} \right]$	$V(x_1, t) = E e^{\gamma x_1/2} \left[ 2 - \gamma x_1 \int_{x_1/c}^t \frac{J_1 \left[ \frac{\gamma c}{2} f_1(\tau) \right] d\tau}{f_1(\tau)} + \frac{\gamma^2 c^2}{4} \left( t - \frac{x_1}{c} \right)^2 \right. \\ \left. - \frac{\gamma^3 c^3 x_1}{8} \int_{x_1/c}^t \frac{(t-\tau)^2}{f_1(\tau)^2} - \gamma c \int_{x_1/c}^t J_0 \left[ \frac{\gamma c}{2} f_1(\tau) \right] d\tau \right] N \left( t - \frac{x_1}{c} \right)$
Short-circuited line of length $x_1$ (Receiving end current before reflections from sending end)	$I(x_1, s) = \frac{2E e^{-\gamma x_1/2}}{L_0 s^2} \left[ \sqrt{\frac{s^2}{c^2} + \frac{\gamma^2}{4}} \epsilon^{-\sqrt{(s^2/c^2) + (\gamma^2/4)x_1}} \right]$	$i(x_1, t) = \frac{2E}{\sqrt{L_0}} e^{-\gamma x_1/2} \left[ \frac{\gamma^2 c t}{4} \int_{x_1/c}^t J_0 \left[ \frac{\gamma c}{2} f_1(\tau) \right] d\tau + J_0 \left[ \frac{\gamma c}{2} f_1(t) \right] \right. \\ \left. - \frac{\gamma^3 c^3}{8} \int_{x_1/c}^t \frac{(t-\tau)^2 J_0 \left[ \frac{\gamma c}{2} f_1(\tau) \right] d\tau}{f_1(\tau)} - \frac{\gamma^2 c^2}{4} \left( t - \frac{x_1}{c} \right) \right] N \left( t - \frac{x_1}{c} \right)$
Line terminated in resist- ance $R$ equal to impedance level at $x=x_1$ (Receiving end voltage before reflections from sending end)	$V(x_1, s) = \frac{E}{s} e^{\gamma x_1/2} \left[ \epsilon^{-\sqrt{(s^2/c^2) + (\gamma^2/4)x_1}} \right] \left[ \frac{2\sqrt{\frac{s^2}{c^2} + \frac{\gamma^2}{4}} R}{L_0 s^2 + R \left( \frac{\gamma}{2} + \sqrt{\frac{s^2}{c^2} + \frac{\gamma^2}{4}} \right)} \right]$	$V(x_1, t) = E e^{\gamma x_1/2} \left[ \frac{\gamma^2 c x_1}{4} \int_{x_1/c}^t \frac{J_1 \left[ \frac{\gamma c}{2} f_1(\tau) \right] d\tau}{f_1(\tau)} - \frac{\gamma c}{2} \left( t - \frac{x_1}{c} \right) - \frac{\gamma c}{2} f_1(t) J_1 \left[ \frac{\gamma c}{2} f_1(t) \right] \right] N \left( t - \frac{x_1}{c} \right) \\ + J_0 \left[ \frac{\gamma c}{2} f_1(t) \right] + \left( \frac{\gamma^2 c^2 t}{4} + \frac{\gamma c}{2} \right) \int_{x_1/c}^t J_0 \left[ \frac{\gamma c}{2} f_1(\tau) \right] d\tau - \frac{\gamma c}{2} f_1(t) J_1 \left[ \frac{\gamma c}{2} f_1(t) \right] N \left( t - \frac{x_1}{c} \right)$

Note:  $f_1(\tau) = \sqrt{\tau^2 - \frac{x_1^2}{c^2}}$  for this and following cases

$$R = \sqrt{\frac{L_0}{C_2}} e^{\gamma x_1}$$

thereafter, increasing in magnitude with time. Fig. 5 shows a curve of the resulting receiving end voltage. Although data on voltage with loads different from the "impedance level" would be of some interest, the transforms involved have proved complex and, to date, uneconomical to evaluate.

The distortion undergone by an ideal rectangular pulse transformed by a section of exponential line may be determined by superposition of two step-voltage responses of opposite polarity with delay between steps equal to the pulse duration. It is apparent from Fig. 5 that distortion may be reduced to any desired extent by reducing the abscissa  $ct/x$  to as low a value as necessary. This may be done, for instance, by choosing a sufficiently low value of the flare coefficient. Flare coefficient is also important, however, in connection with efficiency.

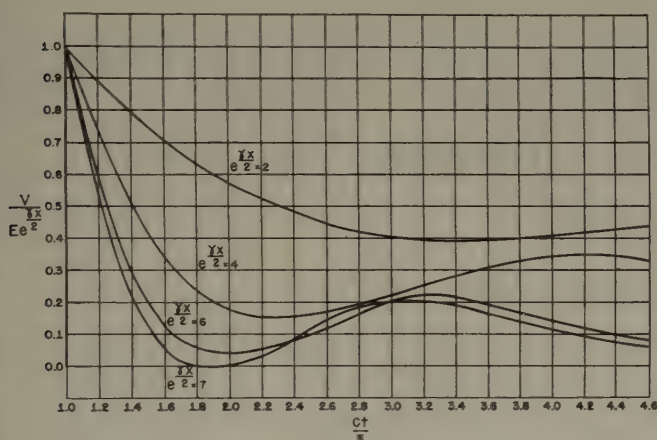


Fig. 5—Receiving voltage as a function of time in terminated exponential transmission-line sections excited with a step input voltage for step-up ratios of one to two, four, six, and seven. The line is terminated in a load resistance equal to its impedance level. Voltage is expressed as the ratio of actual voltage to input voltage multiplied by step-up ratio.

The magnitude of the flare coefficient affects efficiency (a) through its influence on ratio of useful energy delivered to that stored and reflected, and (b) through its effect on dissipation. The first of these effects is significant even when dissipation is negligible. Although all the energy in a dissipationless line is eventually delivered to the load or returned to the generator, the useful load energy is generally only that delivered during the pulse duration. Furthermore, pulse-generator capacity at the sending end is determined by the peak power and energy delivered in a pulse and it is of little advantage to return some of this energy at a later time. It is, therefore, convenient to evaluate a "nominal" efficiency  $n$ , defined as

$$n = \frac{\text{energy delivered to load during duration of initial pulse}}{\text{energy supplied to line at sending end during initial pulse}}$$

Typical nominal efficiency curves are shown in Fig. 6.

The flare coefficient affects dissipation because choice of very low flare coefficient values results in such great

lengths of line to obtain a specified impedance ratio that dissipation may be appreciable. The precise evaluation of dissipation effects on efficiency and pulse shape requires a rigorous transient solution for the exponential line with finite losses; this solution, however, offers extraordinary analytical difficulties. An approximate analysis is readily applicable, however, when losses are not negligible but low.

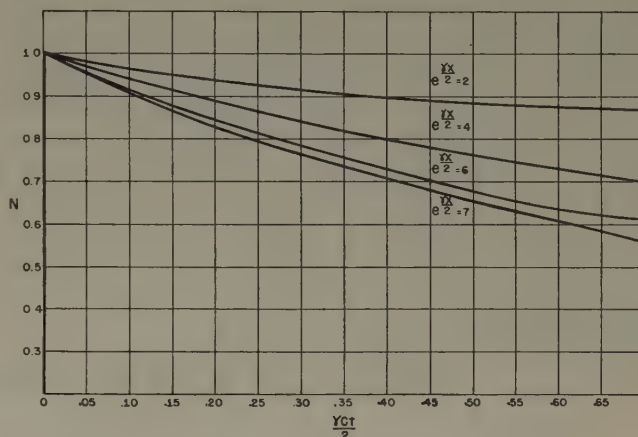


Fig. 6—Nominal efficiency of the exponential-line sections of Fig. 5.

In an analysis of approximate attenuation in uniform lines with losses Ramo and Whinnery<sup>4</sup> have described a method based on the assumption that losses, when small, can be calculated by allowing voltages and currents as computed for a lossless line to encounter the known resistances and conductances. The voltage on a uniform line with losses is of the form

$$V = e^{-\alpha x} f(x, t)$$

in which  $f(x, t)$  is the expression determined for a lossless line.  $\alpha$  is shown to be

$$\alpha = \frac{r}{2Z_0} + \frac{gZ_0}{2}$$

in which  $r$ ,  $g$ , and  $Z_0$  are series resistance per unit length, shunt conductance per unit length, and characteristic impedance, respectively. This expression can be shown to be applicable to exponential lines in a form in which the attenuation per unit length is a function of the parameters associated with that unit length,  $r_x$ ,  $g_x$ , and steady-state characteristic impedance  $Z_{0x}$  at some frequency  $f$ , in terms of which  $\alpha_x = ((r_x/2Z_{0x}) + (g_xZ_{0x}/2))$ . Although  $\alpha_x$  is an attenuation referred to steady-state phenomena, this expression may readily be used to determine the attenuation undergone by the individual frequency components of transient impulses. The components of particular interest are generally the highest frequency components, the attenuation of which is responsible for deterioration of steep leading and trailing

<sup>4</sup> S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, Inc., New York, N. Y., p. 32; 1944.

edges. The frequencies of such components are remote from the steady-state cutoff frequency of a transformer section and the characteristic impedance  $Z_{0x}$  becomes the impedance level  $R_x$ . Burrows<sup>2</sup> has given, in the steady-state case for frequencies remote from cutoff,

$$\alpha_x = \frac{r_x}{2} \sqrt{\frac{C_0}{L_0}} \epsilon^{-\gamma x} + \frac{g_x}{2} \sqrt{\frac{L_0}{C_0}} \epsilon^{\gamma x}.$$

In exponential-line pulse transformers the resistance  $r_x$  is readily held at negligible levels. The conductance  $g$  is likely to be of importance since it is usually convenient to limit the physical length required for a particular minimum pulse distortion by the use of solid or

liquid dielectric material in which there are unavoidable losses. In such cases, the conductance  $g_x$  is an exponential function of length and the attenuation constant independent of  $x$ . For ordinary dielectrics  $\alpha$  is given with sufficient accuracy by

$$\alpha = \frac{\omega}{2} \sqrt{L_0 C_0} \sqrt{\frac{\delta^2}{1 - \delta^2}} = \frac{\omega}{2c'} \sqrt{\frac{\delta^2}{1 - \delta^2}}$$

in which  $\delta$  is the power factor of the dielectric at the frequency  $f$  and  $c'$  the propagation velocity in the dielectric, neglecting losses. Phase distortion owing to changes of velocity with frequency of pulse components is usually of negligible proportions.

## Discussion on

# "On the Energy-Spectrum of an Almost Periodic Succession of Pulses"\*

G. G. MACFARLANE

**T. S. George:**<sup>1</sup> There is an alternate approach to the problem of calculating the power spectra of fluctuating pulses which I have found in general to be more facile than the one employed by Mr. MacFarlane. I am sure that this method is familiar to Mr. MacFarlane, but may not be to other readers. This is the method of first calculating the autocorrelation function, and then transforming to the power spectrum by the Wiener-Khinchine theorem.<sup>2</sup>

Consider, as an example, the first case calculated by Mr. MacFarlane, that is, the case of periodic pulses of shape  $f(t)$  whose amplitude fluctuates in a random fashion described by a probability density function  $P(y)dy$ .

First of all, one may assume that the amplitude of the fluctuating pulses consists of a constant or dc term plus a part which fluctuates about an average value zero. Thus for this part, we have a sequence of periodic pulses with constant amplitude

$$\bar{A} = \int_{-\infty}^{\infty} yP(y)dy.$$

One may proceed directly to the power spectrum through the Fourier series of the pulses. The Fourier series is

$$\sum_{-\infty}^{\infty} c_n e^{j2\pi n t/T}.$$

The power spectrum then is

$$2\pi \sum_{-\infty}^{\infty} |c_n|^2 \delta\left(\omega - \frac{2\pi n}{T}\right)$$

where

$$c_n = \frac{\bar{A}}{T} \int_0^T f(t) e^{-2\pi n t/T} dt = \frac{\bar{A}}{T} G\left(\frac{2\pi n}{T}\right)$$

and  $G(w)$  is the continuous spectrum of a single pulse of unit amplitude.  $\delta(\omega)$  is the Dirac delta function, and is defined here as

$$\delta(\omega) = \frac{1}{2\pi} \int_{-\infty}^{\infty} e^{-j\omega t} dt.$$

This accounts for the second part of Mr. MacFarlane's result.

For the fluctuating part, we calculate the autocorrelation  $\rho(\tau) = \overline{f(t)f(t+\tau)}$ . It is observed first that if  $t$  is within one pulse and  $t+\tau$  in any other pulse

$$\overline{f(t)f(t+\tau)} = \overline{f(t)} \overline{f(t+\tau)} = 0,$$

since the separate pulses are assumed to fluctuate independently (tacitly assumed by Mr. MacFarlane). Thus permissible values of  $\tau$  lie between  $-T$  and  $T$  and it is obvious that this part of the spectrum will be continuous. Now it is convenient to consider the fluctuating pulse to consist of  $f(t)$  itself, multiplied by an infinite line parallel to the  $t$  axis and which fluctuates about zero with probability determined by  $P(y)dy$ .

The autocorrelation function of the line is  $\bar{f}^2$ , say,

\* PROC. I.R.E., vol. 37, pp. 1139-1144; October, 1949.

<sup>1</sup> Philco Corporation, Philadelphia, Pa.

<sup>2</sup> H. M. James, N. B. Nichols, R. S. Philips, "Theory of Servomechanisms," Radiation Laboratory Series No. 25, p. 283, McGraw-Hill Book Co., New York, N. Y.; 1947.

where  $I$  is the instantaneous departure of the line from zero. This is, of course,

$$(A - \bar{A})^2.$$

The power spectrum of this line is obtained from  $\rho(\tau)$  by the Wiener-Khintchine theorem which for unnormalized  $\rho(\tau)$  may be written

$$F_1(\omega) = \int_{-\infty}^{\infty} \rho(\tau) \cos \omega \tau d\tau.$$

Thus for the line,

$$F_1(\omega) = 2\pi(A - \bar{A})^2 \delta(\omega).$$

The average power spectrum of the pulse itself is

$$\frac{|G(\omega)|^2}{T}.$$

Now when one has the product of two time functions, whose spectra are  $F_1(\omega)$  and  $F_2(\omega)$ , respectively, the complex convolution integral gives the spectrum of the product, thus

$$F(\omega) = \frac{1}{2\pi} \int_{-\infty}^{\infty} F_1(\omega - s) F_2(s) ds.$$

Thus the continuous power spectrum is

$$\begin{aligned} F(\omega) &= \frac{2\pi(A - \bar{A})^2}{2\pi} \int_{-\infty}^{\infty} \frac{|G(\omega - s)|^2}{T} \delta(s) ds \\ &= \frac{|G(\omega)|^2}{T} (A - \bar{A})^2. \end{aligned}$$

This is the other part of Mr. MacFarlane's result.

Several steps are involved in this method, but each is easy to carry out. The same procedure may be applied to the problem of pulses varying statistically in width or position.

**G. G. Macfarlane:**<sup>3</sup> I agree with Mr. George that the results given in equations (21) and (33) can be more neatly derived with the aid of Khintchine's theorem. Moreover, the forms of the autocorrelation functions are themselves of interest. Nevertheless, the method given in the paper was adopted because it seems to the author to be more elementary.

<sup>3</sup> Ministry of Supply, Great Malvern, Worcs., England.

## Measurement of the Electrical Characteristics of Quartz Crystal Units by Use of a Bridged-Tee Network\*

CHARLES H. ROTHAUGE†, ASSOCIATE, IRE, AND FERDINAND HAMBURGER, JR.‡, SENIOR MEMBER, IRE

**Summary**—A bridged-tee null network has been used to measure the equivalent resistance and the equivalent reactance of a quartz crystal plate.

This measuring circuit has the advantages that shielding is relatively simple (both the source and the detector have a common grounded terminal) and that corrections for all stray capacitances that affect the measurements may be included in the calibrations of the capacitors of the tee.

The precision of the measurements is estimated to be 0.3 per cent for the determination of the equivalent reactance and 2.3 per cent for the determination of the equivalent resistance.

### INTRODUCTION

WITH THE EXTENDED USE of quartz crystal plates in the past several years there has been an increasing demand for the accurate measurement of the electrical properties of the crystals. In the past, the expression of the dynamic impedance of the crystal has been made in an arbitrary and indirect man-

ner.<sup>1,2</sup> However, the electrical characteristics (the impedance, the equivalent reactance, and the equivalent resistance) of a quartz crystal plate may be measured at any frequency of operation without reference to the external circuit in which the unit is to be used.<sup>3-7</sup> These electrical quantities permit the prediction of the performance of the crystal unit in any particular external circuit whose characteristics are known. Such a method of specification of crystal units using their impedance,

<sup>1</sup> G. M. Thurston, "A crystal test set," *Bell Lab. Rec.*, vol. 22, pp. 477-480; August, 1944.

<sup>2</sup> W. E. McNatt, "Test set for quartz crystals," *Electronics*, vol. 18, pp. 113-115; April, 1945.

<sup>3</sup> C. W. Harrison, "The measurement of the performance index of quartz plates," *Bell Sys. Tech. Jour.*, vol. 24, pp. 217-252; April, 1945.

<sup>4</sup> A. J. Biggs and G. M. Wells, "The measurement of the activity of quartz oscillator crystals," Part III, *Jour. IEE* (London), vol. 93, pp. 29-36; January, 1946.

<sup>5</sup> K. S. Van Dyke and A. M. Thorndyke, "The three-crystal method of quartz resonator measurement," *Phys. Rev.*, vol. 57, p. 560; March, 1940.

<sup>6</sup> C. J. Miller, Jr., "Long Range Planning for Quartz Crystal Units," Signal Corps Engineering Laboratories; October, 1945.

<sup>7</sup> W. D. George, M. C. Selby, and R. Scolnik, "Precision measurement of the electrical characteristics of quartz crystal units," *PROC. I.R.E.*, vol. 36, pp. 1122-1131; September, 1948.

\* Decimal classification: R214.211. Original manuscript received by the Institute, November 26, 1949; revised manuscript received, April 21, 1950.

† U. S. Naval Postgraduate School, Annapolis, Md.

‡ The Johns Hopkins University, Baltimore, Md.

equivalent reactance, and resistance has many advantages,<sup>3,8</sup> not the least of which is the availability to the engineer of data which will be most useful in design procedures.

This paper will present a method of measurement of these electrical characteristics of a quartz crystal unit using a "bridged-tee" null network.

#### EQUIVALENT CIRCUIT OF A PIEZOELECTRIC CRYSTAL

A piezoelectric crystal vibrating near its resonant frequency may be represented by an equivalent electric circuit.<sup>9</sup> This equivalent circuit, composed of a series branch of resistance, inductance, and capacitance in parallel with a capacitance, may be reduced to a series arrangement of an effective inductance and an effective resistance when the frequency of operation of the crystal is between its series resonant and antiresonant frequencies. The values of the series resistance and inductance vary rapidly as the frequency is changed from series to parallel resonance. This simplified circuit is the one that will be used to represent a crystal in this paper and implies that the crystal resonators under consideration are to be operated into capacitive loads.

#### THE BRIDGED-TEE NULL NETWORK

A tee configuration of two equal capacitances and a resistance will form a null network when bridged by an inductance in series with a resistance.<sup>10</sup> This arrangement is shown in Fig. 1.

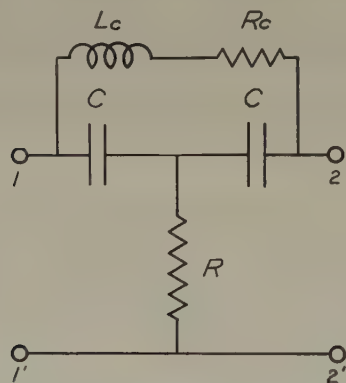


Fig. 1—Bridged-tee null network.

This circuit may be used to measure the effective inductance and resistance of a crystal at its operating frequency, when the crystal is one designed to operate into a capacitive load.

The balance conditions for such a network (i.e., conditions for zero transmission) are

$$X_C = \frac{2}{WC} \quad (1)$$

<sup>8</sup> K. S. Van Dyke, "The standardization of quartz crystal units," *Proc. I.R.E.*, vol. 33, pp. 15-20; January, 1945.

<sup>9</sup> K. S. Van Dyke, "The electrical network equivalent of a piezoelectric resonator," *Phys. Rev.*, vol. 25, p. 895; June, 1925.

<sup>10</sup> W. N. Tuttle, "Bridged-T and parallel-T null circuits for measurements at radio frequencies," *Proc. I.R.E.*, vol. 28, pp. 23-29; January, 1940.

$$R_C = \frac{1}{RW^2C^2} \quad (2)$$

Such an arrangement has several advantages which are readily apparent. First, stray capacitances between the terminals 1 and 1' and terminals 2 and 2' do not affect the measurement. Effectively such capacitances are across the source and the detector. Second, the source and detector have a common grounded terminal. Shielding is thus simplified since no shielded transformers or Wagner ground connections are needed.

Stray capacitance from the junction point of the tee to ground will introduce errors. The conditions for zero transmission with such a stray capacitance  $C_1$  become

$$R_C = \frac{1}{RW^2C^2} \quad (3)$$

$$X_C = \frac{2}{WC} \left( \frac{C_1}{2C} + 1 \right) \quad (4)$$

#### CIRCUIT ARRANGEMENTS AND COMPONENTS

The circuit arrangement consists of the source, two buffer amplifiers, the bridged tee, and a detector. These units and the shielding are shown in the schematic diagram of Fig. 2 and will be briefly described.

The circuit was driven by a crystal controlled triode oscillator (source). This oscillator is controlled by a crystal having approximately the same frequency as the crystal being measured. An adequate range of frequencies was obtained by use of a small variable air capacitor placed across the crystal. Additional range may be obtained by the use of a series inductance and by a suitable selection of crystals with slightly different natural resonant frequencies. The frequency stability of the oscillator was better than one part in  $10^7$  for the periods during which measurements were taken.

It was found necessary to isolate the bridged tee from the remainder of the circuit in order to prevent coupling between the source and the detector. Isolation was accomplished by use of two buffer amplifier stages, a triode amplifier coupled inductively to the source and capacitively to the tee and a miniature pentode amplifier coupled capacitively to the output of the tee.

The triode amplifier is of conventional resistance-coupled design using a 6AC5 high-mu power amplifier triode. The gain of this stage is approximately 12 to 15 for the frequency range of 7 megacycles per second to 1 megacycle per second. The buffer amplifier using a 6AK5 pentode was designed with a tuned plate circuit and an input of the highest possible impedance. Its gain was approximately 25 at 7 megacycles per second. The high input impedance increases the sensitivity of the bridged-tee network.

A radio receiver was used as a detector for the measurements. The output of the buffer amplifier was connected directly to the antenna input of the receiver, and the input signal strength determined by measuring the

## MEASUREMENTS

Figure 1 is a schematic diagram of a three-stage amplifier. It consists of three main sections: a Source and Buffer Amplifier, a Bridged Tee, and a Buffer Amplifier. The Source and Buffer Amplifier uses a 6J5 tube and includes a 10K resistor, a 100 nF capacitor, and a 25K load resistor. The Bridged Tee section uses a 6AC5 tube and features a 10K resistor, a 100 nF capacitor, and a 25K load resistor. The Buffer Amplifier section uses a 6AK5 tube and includes a 10K resistor, a 100 nF capacitor, and a 25K load resistor. The output of the Buffer Amplifier is connected to a speaker. The entire circuit is powered by a B+ (250V) supply.

Fig. 2—Schematic diagram of measuring circuit.

minimum output is obtained; then the resistive balance is made with the resistor  $R$ .

Inspection of the balance conditions (3) and (4) reveals why this order of operation is essential. Both the reactive and resistive balance depend upon the value of capacitance, while the variation of  $R$  affects only the resistive balance. In addition, variations of detector voltage caused by variations of  $C$  are many times that caused by similar variations in  $R$ .

Thus, after an approximate reactive balance is obtained, the gain of the detector may be increased to secure the desired sensitivity to determine the resistive balance and the final fine adjustment of capacity and resistance. The gain of the detector must not be placed too high until after a coarse reactive balance is obtained, since the relatively large input will mask any indication of a minimum signal.

The limits of the bridged tee are dependent upon frequency. With the values of  $C$  used in the circuit reported in this paper, inductive reactances of 100 to about 1,520 ohms may be measured at a frequency of 7 megacycles per second, reactances of 700 to 10,600 ohms at one megacycle per second. These limits may be extended but it is to be noted that a practical limit is reached for the larger values of reactance when the stray capacitance becomes appreciable in relation to the value of  $C$ . No difficulty is encountered in lowering the limits.

The limits of the value of resistance that may be determined are dependent not only on the value of  $R$  but also upon the reactive balance. If the inductive reactance of a series combination of resistance and reactance is 100 ohms at a frequency of 7 megacycles per second, the resistance value that can be determined is between 0.15 and 300 ohms; at the same frequency, with an inductive reactance of 1,520 ohms, resistances of 40 to 55,000 ohms may be determined. Similarly, at one megacycle per second, resistance values of 8 to 12,000 ohms may be measured when associated with an inductive reactance of 700 ohms, with corresponding values of resistances with an inductive reactance of 10,600 ohms.

Several series of crystals have been measured. The detailed data will not be given in this paper. However, representative results are given in Table I. The nominal frequency of all crystals in a series was the same, and the measurements reported were made at the same frequency for each series.

Values of  $X_C$  were measured from about 218 ohms

to 747 ohms, and values of  $R_C$  ranged from approximately 10 ohms to 33 ohms.

TABLE I  
CRYSTAL TYPE DC-11-A<sup>1</sup>

Crystal No.	$X_C$ (ohms)	$R_C$ (ohms)
1	522.0	32.19
2	415.0	30.50

CRYSTAL TYPE DC-31-C.P.R.<sup>2</sup>

Crystal No.	$X_C$ (ohms)	$R_C$ (ohms)
1	662.0	12.22
2	650.2	18.46
4	707.0	13.72
8	594.0	22.80

<sup>1</sup> Measurements made at a frequency of 7,102,115 cycles per second.

<sup>2</sup> Measurements made at a frequency of 4,540,362 cycles per second.

The precision of these measurements is estimated to  $\pm 0.3$  per cent for the determination of  $X_C$  and  $\pm 2.3$  per cent for the determination of  $R_C$ . This estimate was made from a series of ten measurements made at the same frequency on the same crystal. The average value of  $X_C$  was 532.9 ohms with an average deviation from the mean of  $\pm 1.8$  ohms or  $\pm 0.3$  per cent. The average value of  $R_C$  was 13.03 ohms with an average deviation from the mean of  $\pm 0.30$  ohms or  $\pm 2.3$  per cent. The measurements were made at a frequency of 4,950,340 cycles per second.

## CONCLUSIONS

A method of measurement is presented for the determination of the equivalent series reactance and the equivalent series resistance of a quartz crystal plate at its operating frequency. This method utilizes a bridged-tee null network.

The determination of the equivalent series reactance is made with a precision of  $\pm 0.3$  per cent. The equivalent series resistance is determined with a precision of  $\pm 2.3$  per cent.

A design of a variable resistance box is presented. The box has values of resistance of 10 to 15,210 ohms in steps of 10 ohms. The resistance values of the box remain within 5 per cent of their direct-current values up to frequencies of 7 megacycles per second with an equivalent capacitance to its shield of  $23.5 \pm 0.3$  micromicrofarads for all values of resistance.

## CORRECTION

George Sinclair, author of the paper, "The transmission and reception of elliptically polarized waves," which appeared on pages 148-151 of the February, 1950, issue of the PROCEEDINGS OF THE I.R.E., has brought the following error to the attention of the editors.

On page 149, the vector  $N$  in equations (2) and (3) should read  $N_t$ , where  $N_t$  is the component of  $N$  transverse to the direction of propagation.

# Contributors to Proceedings of the I.R.E.

Alfred C. Beck (A'30-SM'46) was born on July 26, 1905, at Granville, N. Y. He received the E.E. degree from Rensselaer

Polytechnic Institute in 1927. After two summers in the test department of the New York Edison Company and a year as instructor in mathematics at Rensselaer, he became a member of the technical staff of Bell Telephone Laboratories, Inc., in 1928. Since then he has been in

the radio research department, working on antennas, waveguides, and various short-wave, radar, and microwave projects. He is a New York State licensed professional engineer and a member of Sigma Xi.



Thomas P. Cheatham, Jr., was born on January 30, 1923, in Washington, D. C. After receiving the B.S. degree and a commission as ensign from the United States Coast Guard Academy in June, 1943, he served three

years of active duty, leaving the service with the rank of lieutenant. He entered Massachusetts Institute of Technology in June, 1946, to pursue graduate study and research, and received the M.S. degree in electrical engineering in September, 1947. He was a research associate in the Research Laboratory of Electronics from 1946 to 1949.

Mr. Cheatham is a member of Sigma Xi. He is now associated with the Optical Research Laboratory and physics department of Boston University.



Takeo Hada was born on November 7, 1920, at Osaka, Japan. He was graduated from the electrical engineering department of the Osaka High Technical School in 1939, and then entered the Kawanishi Machine Works.

Mr. Hada is now associated with the Kobe Kogyo Corporation, in Kobe, Japan, engaged in research.

He is a member of the Institute of Electrical Engineering and Electrical Communication in Japan.

Robert W. Dawson was born in Spring Lake, N. J., on May 26, 1922. He joined the radio research department of the Bell

Telephone Laboratories, Inc., in 1941, after attending Newark College of Engineering. Mr. Dawson enlisted in the U. S. Army Signal Corps in 1942. He received training as a wire chief at Ft. Monmouth, and was then selected to attend the Army Specialized Training Pro-

gram at Rutgers University, where he completed the course in communications engineering in 1944. Assignment was then made to the Manhattan District at Los Alamos, N. Mex., where he remained until his discharge, except for a few months overseas with the Atom Bomb group from Los Alamos. A special citation was received from Rear Admiral Purnell USN, for this overseas work.

After discharge from the Army early in 1946, employment was resumed with the Bell Telephone Laboratories at Holmdel, N. J., where Mr. Dawson now works on various microwave projects.



Albert A. Gerlach (S'42-A'46) was born on May 22, 1920, in Columbus, Ohio. He received the B.S. degree in electrical engineering from the Ohio State University in June, 1942, and the M.S. degree in electrical engineering from the Illinois Institute of Technology in January, 1949.

From 1942 to 1946, as a radar officer in the U. S. Army Signal Corps, he instructed and supervised the instruction of military personnel in the theory, operation, and maintenance of radar equipments, wrote technical manuals on radar and radio equipments, and made siting surveys on the West Coast for purposes of radar air-warning installations. In 1946, he became senior project engineer and, subsequently, laboratory director at the Rowe Engineering Corporation in Chicago and, since 1948, he has been a research engineer at the Armour Research Foundation of the Illinois Institute of Technology.

Mr. Gerlach is the vice-chairman of the Procedures Committee of the Chicago Section of the IRE, an associate member of AIEE and Sigma Xi, and a member of Eta Kappa Nu and Tau Beta Pi. He is a registered professional engineer in the State of Illinois.

F. Hamburger, Jr., (A'32-M'39-SM'43) was born in Baltimore, Md., on July 5, 1904. He received the B.E. degree in electrical engineering from The Johns Hopkins University in 1924. After participating in a program of dielectric research for several years, he earned the degree of Doctor of Engineering from that University in 1931.

Since 1931, Dr. Hamburger, has been on the staff of the electrical engineering department of The Johns Hopkins University where he was appointed professor of electrical engineering in 1947. He served as chief test engineer for Bendix Radio Division from 1942 to 1945 while on partial leave of absence from the University, as consultant for the National Defense Research Council, Section 17-2, during 1944-1945, and as consultant to the Research and Standards Section, Bureau of Ships, Navy Department, during 1945-1946. At the present time he is associate director in charge of engineering of the systems research contract between The Johns Hopkins University and Special Devices Center, Office of Naval Research.

Dr. Hamburger has served as chairman of the Baltimore Section of the IRE in 1940-1941 and is now regional director. He is also a fellow of the American Institute of Electrical Engineers, a member of Sigma Xi and Tau Beta Pi, and a director of The Engineers Club of Baltimore.



John V. Harrington (A'47) was born in New York, N. Y., on May 9, 1919. He received the B.E.E. degree from Cooper Union

Institute of Technology in 1940 and the M.E.E. degree from Polytechnic Institute of Brooklyn in 1948. At the present time he is a special graduate student at the Massachusetts Institute of Technology.

Upon completion of his undergraduate work, he spent two years in the power engineering field, first as a student engineer with the Consolidated Edison Company of New York and then as an assistant engineer with the American Gas and Electric Service Corporation. At the beginning of the war he spent several months as a research associate working on search receivers at the Radio Research Laboratory of Harvard University, and then accepted a commission in the Navy.



A. C. BECK



R. W. DAWSON



F. HAMBURGER, JR.



T. CHEATHAM, JR.



A. A. GERLACH



J. V. HARRINGTON



TAKEO HADA

While a member of the Armed Forces, Mr. Harrington served as an electronic specialist in both the United States and the Pacific theatres. Following separation in 1946, he joined the staff of the Air Force Cambridge Research Laboratories where he is a member of the Communications and Relay Laboratory and has been engaged in research in the fields of microwave relay, data transmission, and storage systems.

Mr. Harrington is a member of Tau Beta Pi and Sigma Xi.



Robert L. Henry (A'50) was born on March 13, 1918, in Rutledge, Tenn. He received the B.S. degree in mechanical engineering in 1939 from Clemson College.



ROBERT L. HENRY

After two years of employment in machine design and production engineering positions, he was called to active duty as a captain with the Army Ordnance Department. The last two years of army service were with the Office of the Chief of Ordnance in connection with the proximity-fuze program.

Upon discharge from the service in 1946 he joined the Ordnance Development Division of the National Bureau of Standards, where he soon became associated with the Bureau's activity in the field of printed electronic circuits. Mr. Henry is at present employed in the Electronics Division of the National Bureau of Standards, conducting research on materials and techniques associated with special production processes and the mechanization of the production of electronics equipment. He is a member of Tau Beta Pi.



Y. W. Lee was born on April 14, 1904, at Macao, China. He received the Sc.D. degree in electrical engineering in 1930 from the Massachusetts Institute of Technology.



Y. W. LEE

Dr. Lee was a professor of electrical engineering at National Tsing Hua University, Peiping, China from 1934 to 1937, at St. John's University from 1942 to 1946, and at Ta Tung University, in Shanghai, China, during the same period.

At present, he is an associate professor of electrical engineering at Massachusetts Institute of Technology.

B. Jennings (S'38) was born on September 13, 1910, at Baltimore, Md. He attended The Johns Hopkins University as a graduate student, specializing in nuclear physics up to the beginning of the war, and received the M.A. degree in 1947.



B. JENNINGS

During the war he became a test equipment design engineer at Western Electric Company at Baltimore, Md. In 1946 he joined the electronics and nuclear physics department of the Westinghouse Research Laboratories and was engaged in radar design. He became one of the members of the nuclear physics group in 1947 which was being reformed after the war, and in 1949 was made section manager in charge of nuclear research at the Westinghouse Research Laboratories, his present position.



Tadashi Nakamura was born on December 23, 1923, at Ujiyamada, Japan. He was graduated from the electrical engineering department of Nagoya Technical College in 1943, and joined the technical staff of the Kawasaki Machine Works.



T. NAKAMURA

At present Mr. Nakamura is affiliated with the Kobe Kogyo Corporation, in Kobe, Japan.

He is a member of the Institute of Electrical Engineering and Electrical Communication in Japan.



Ken-ichi Owaki was born on August 25, 1910, at Kochi, Japan. He was graduated from the electrical engineering department of Nagoya Technical College in 1934, and received the Ph.D. degree from Osaka University in 1948.

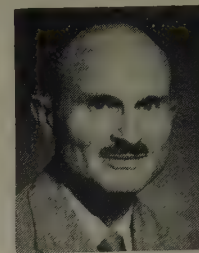


KEN-ICHI OWAKI

From 1935 to 1937 Dr. Owaki was an assistant in the science department of Osaka University, until he entered the Kawasaki Machine Works in 1937.

He is a member of the Institute of Electrical Engineering and Electrical Communication in Japan. He is now associated with the Kobe Kogyo Corporation in Kobe, Japan.

Ronold King (A'30-SM'43) was born on September 19, 1905, at Williamstown, Mass. He received the B.A. degree in 1927 and the



RONOLD KING

M.S. degree in 1929 from the University of Rochester, and the Ph.D. degree from the University of Wisconsin in 1932. He was an American-German exchange student at Munich from 1928 to 1929; a White Fellow in physics at Cornell University from 1929 to 1930; and a Fellow in electrical engineering at the University of Wisconsin from 1930 to 1932. He continued at Wisconsin as a research assistant from 1932 to 1934. From 1934 to 1936 he was an instructor in physics at Lafayette College, serving as an assistant professor in 1937.

During 1937 and 1938, Dr. King was a Guggenheim Fellow at Berlin. In 1938 he became instructor in physics and communication engineering at Harvard University, advancing to assistant professor in 1939 and to associate professor in 1942. He was appointed Gordon McKay professor of applied physics at Harvard University in 1946.

Dr. King is a Fellow in the American Physical Society, the American Association for the Advancement of Science, and the American Academy of Arts and Sciences. He is a member of Phi Beta Kappa and Sigma Xi.



William G. Pfann was born in New York, N. Y., on October 25, 1917. He received the B.Ch.E. degree from Cooper Union in



W. G. PFANN

1940. He is now a research metallurgist on the technical staff of the Bell Telephone Laboratories, Inc., where he has been employed since 1935.

Mr. Pfann has been engaged principally in research on semiconductors for crystal rectifiers and transistors.

He is a member of

Tau Beta Pi, the American Physical Society, and the American Institute of Mining and Metallurgical Engineers.



Thomas F. Rogers (M'50) was born in Providence, R. I., on August 11, 1923. He received the B.S. degree in physics from Providence College in January, 1945, and the M.A. degree, also in physics, from Boston University in August, 1949.

From January, 1945, until the end of the

war, he was on the staff of the Radio Research Laboratory of Harvard University, where, as a research associate, he was engaged in radar-countermeasures work. For the next year he performed circuit design on projection-type television receivers at the Bell and Howell Company in Chicago, Ill. During much of this time he was also a radio instructor at the Chicago Technical College.



T. F. ROGERS

Mr. Rogers joined the staff of the Air Force Cambridge Research Laboratories in Cambridge, Mass., in August, 1946, and has since been associated with this organization. As a member of the Communications and Relay Laboratory at Cambridge he has engaged in research work in the fields of broad-band microwave relaying, microwave propagation, memory devices, and magneto-acoustic effects. He is a member of the Research and Development Board Subpanel on Special Tubes.



Charles H. Rothauge was born on September 8, 1919, in Baltimore, Md. He received the B.E. degree from The Johns Hopkins University, in 1940, and the Doctor of Engineering degree from the same University in 1949.



C. H. ROTHAUGE

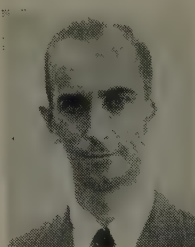
From August, 1941, to April, 1946, he served with the Ordnance Department of the United States Army at Aberdeen Proving Grounds, Md. During this time he conducted tests on aircraft armament. In the spring of 1944, he headed a special test group which conducted high-altitude tests of ordnance material in the Chilean Andes. He was separated from active duty with the rank of major.

During the period from April, 1946, to June, 1949, he was an instructor of electrical engineering at The Johns Hopkins University, and in the summer of 1949 he joined the faculty of the United States Naval Postgraduate School as assistant professor of electrical engineering. Dr. Rothauge is a member of Sigma Xi and Tau Beta Pi.



V. H. Rumsey was born in 1919, in Devizes, England. He graduated from Cambridge in 1941 with an honors degree with distinction in Part III of mathematical tripos. In the United Kingdom civil service, he was a third-class assistant in 1941, a junior scientific officer in 1942, and a scientific officer in 1943 at TRE, Great Malvern, England. From 1943 to 1945 he was the head of the antenna section of the Combined Research Group at the Naval Research Laboratories in Washington, D. C. As a senior scientific officer in the

United Kingdom civil service from 1945 to 1948, he worked as a theoretical physicist at the Canadian atomic energy project. Since 1948, Mr. Rumsey has been supervisor of the Antenna Laboratory at Ohio State University.



V. H. RUMSEY

Mr. Rumsey is a member of the IRE Committee on Antennas and Waveguides, chairman of the Antenna Instrumentation Subpanel of the Research Development Board panel on antennas and propagation, and a member of the URSI #1 Commission on Radio Standards and Measurements.



J. H. Scaff was born in Jackson, Tenn., on October 1, 1908. He received the B.S. degree in chemical engineering from the University of Michigan in 1929. Since then he has been associated with Bell Telephone Laboratories, Inc., as a member of their metallurgical research group.



J. H. SCAFF

Mr. Scaff's work has been principally in the field of semiconductors for use in rectifiers and transistors and magnetic materials. He is an active participant in the affairs of the Institute of Metals Division of AIME as a member of their Executive Committee and as Chairman of their Membership Committee.



Robert K-F Scal (S'41-A'45-M'47) was born on December 8, 1919, in New York, N. Y. After receiving the B.A. degree in chemistry from Stanford University in 1941, he began graduate work in electrical engineering. These studies were interrupted in 1942, when he accepted a commission in the United States Army Signal Corps.



ROBERT K-F SCAL

Following ES-MWT radar courses at Harvard and the Massachusetts Institute of Technology, Mr. Scal transferred to the Army Air Force Air Materiel Command in 1943. Here he was primarily concerned with electronic maintenance, research, and development.

In 1946 Mr. Scal returned to Stanford University to continue his studies. While at Stanford he designed and built new control circuits for the Stanford ionosphere equipment. He received the M.A. degree in electrical engineering in 1946, and the professional degree of Electrical Engineer in 1947.

Since 1947 Mr. Scal has been associated with the Electronics Division of the National Bureau of Standards, where he has been working on miniaturization of electronic equipment and components, and on development of new techniques and materials for production of such equipment.



Edward R. Schatz (A'44) was born in St. Marys, Pa., in 1921. He received the B.S. degree in 1942 and the M.S. degree in 1943, both from the Carnegie Institute of Technology.



EDWARD R. SCHATZ

From 1943 to 1946 he was employed by the Manhattan District Project at the University of Chicago and Los Alamos Scientific Laboratory. He returned to Carnegie Tech in 1946 as an instructor in electrical engineering, and completed the work for the Doctor of Science degree in 1949.

Dr. Schatz is now assistant professor of electrical engineering and industrial administration at Carnegie Tech. He is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.



Gustave Shapiro (A'43) was born in New York, N. Y., on July 6, 1917. He attended George Washington University. Following two years of employment in test and development work, he became a project engineer at the Eatontown Signal Laboratory in 1942. In 1945 he joined the Evans Signal Laboratory, where he worked on direction-finding antenna and miniaturization problems.



GUSTAVE SHAPIRO

From 1946 to 1947 he was engaged in the development of intermediate-frequency amplifiers for moon radar equipment at the Coles Signal Laboratory.

Since 1947 Mr. Shapiro has been on the staff of the Electronics Division of the National Bureau of Standards, where he works on the development of unclassified miniaturization techniques. He is a member of the Research and Development Board Subpanel on Miniature Components and Packaged Sub-Assemblies.

Shigeo Terahata was born on July 14, 1917, at Tokyo, Japan. He received the B.E. degree in 1941 from Tokyo University, and then entered the Kawanishi Machine Works. Mr. Terahata is now employed by the Kobe Kogyo Corporation in Kobe, Japan.



SHIGEO TERAHATA

He is a member of the Institute of Electrical Engineering and Electrical Communication in Japan.



Lester C. Van Atta (M'42-SM'43) was born in Portland, Ore., on April 18, 1905. He received the B.A. degree in 1927 from Reed College; the M.S. degree in 1929, and the Ph.D. degree in physics in 1931 from Washington University in St. Louis, Mo.



L. C. VAN ATTA

Dr. Van Atta conducted research at Washington University, Princeton University, and also at the Radiation Laboratory and in the physics department of the Massachusetts Institute of Technology. He was head of the Antenna Research Branch of the Naval Research Laboratory, from 1945 to 1950. His present position is head of the Antenna Laboratory and member of the Advisory Council of the Hughes Aircraft Company Research and Development Laboratories in Culver City, Calif.

Dr. Van Atta is past Chairman of the Panel on Antennas and Propagation of the Research and Development Board. He served on the following IRE committees: Research, 1946-1947, 1949-1950; Antennas and Wave Guides, Chairman 1949-1950; and the Standards, 1949-1950. He has also served as Chairman of the Administrative Committee of the IRE Professional Group on Antennas and Propagation. He is currently Chairman of the United States Commission 6 of the International Scientific Radio Union. Dr. Van Atta is a Fellow of the Physical Society.

His special fields of interest have included high resistors, electron scattering in gases, high-voltage generation, nuclear reactions by bombardment, and the microwave antenna theory and design.

Jerome B. Wiesner (S'36-A'40-SM'48) was born in Detroit, Mich., on May 30, 1915. He received the B.S. degree in 1937, the M.S. degree in 1940, and the Ph.D. degree in May, 1950, all from the University of Michigan.



J. B. WIESNER

He was chief engineer of the Acoustical and Record Laboratory of the Library of Congress from 1940 to 1942, at which time he became a member of the staff of the Radiation Laboratory at the Massachusetts Institute of Technology. In 1945, Dr. Wiesner went to the Los Alamos Laboratory in New Mexico as a member of the staff, returning to MIT in 1946. He has been assistant director of the Research Laboratory of Electronics at MIT for the past three years and is now associate director. In addition to his work in the laboratory, he is professor of electrical engineering at MIT.

Dr. Wiesner is a member of the Acoustical Society of America, the Federation of American Scientists, and the American Association for the Advancement of Science. He is also a member of Eta Kappa Nu, Sigma Xi, and Phi Kappa Phi.



Everard M. Williams (S'36-A'41-SM'44) was born in New Haven, Conn., in 1915. He received the B.E. degree in 1936 and the Ph.D. degree in 1939, both from Yale University.



E. M. WILLIAMS

During the summer of 1937 he was employed by the General Electric Company, and during the academic year 1938-1939, he was the recipient of a Charles A. Coffin Fellowship from this company. From 1939 to 1942, he was an instructor in electrical engineering at the Pennsylvania State College. From 1942 to 1945, he served as the chief engineer of the development branch, Special Projects Laboratory, Radio and Radar Subdivision, ATSC, Wright Field, Ohio. He subsequently

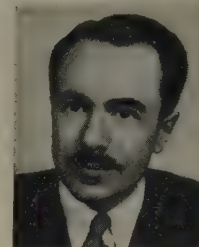
received the President's Certificate of Merit for outstanding service in this post.

In 1945, Dr. Williams was appointed associate professor of electrical engineering at the Carnegie Institute of Technology, in Pittsburgh, Pa., where he is now a professor. He is also serving as expert consultant to the Research and Development Board of the National Military Establishment.

Dr. Williams was the recipient of the 1946 Eta Kappa Nu award. During 1947-1948 he served as Chairman of the IRE Pittsburgh Section.



Raymond M. Wilmotte was born in Paris, France, on August 13, 1901. He took first class honors in Mechanical Science (Engineering) at Cambridge University, England, in 1920, when he received the M.A. degree.



R. M. WILMOTTE

Until 1929 he was engaged at the National Physical Laboratory in research on radio measurements, direction finding, wave propagation, and theory and design of antennas.

In 1929 Mr. Wilmotte came to the United States and worked until 1931 at Boonton Aircraft Corporation on research and development on aircraft equipment, including antennas and blind landing systems. He opened a consulting practice in 1931 on being requested by Commander Craven to solve the problem of a client of his, Station WFLA, which interfered with Station WTMJ. This led him to design, install, and test the first directional antenna used for this purpose. He has appeared in patent cases and before the Federal Communications Commission. During the war he carried out design and development contracts for the Signal Corps, the Navy, and NDRC. At the request of the latter he established a small manufacturing facility for specialized production where one of the projects was the modification of radar units to meet the Japanese Kamikaze menace. For this work he was awarded the Bureau of Ordnance Development Award.

Mr. Wilmotte invented the antifading broadcast antenna recently used in Germany. Recently he has demonstrated the possibility of receiving a weak FM signal in the presence of a strong one in the same frequency band. He was one of the original members of the "Ad Hoc" Committee of the Federal Communications Commission.



# Correspondence

## Propagation of UHF and SHF Waves Beyond the Horizon\*

Calculations of ground-wave propagation beyond the horizon by the usual methods are severely limited at frequencies above 200-500 megacycles by the uncertainty in the choice of effective antenna heights, and the clearance or lack of clearance over some intervening obstacle is frequently more important than the antenna heights at the terminals. It now appears that a useful parameter is the actual clearance divided by the height of the first Fresnel zone. The diffraction loss over a perfect sphere, as well as diffraction over a knife edge and reflection from a plane surface, can be expressed in terms of this ratio, and the results are illustrated on Fig. 1.

The plane earth reflection theory shows the familiar series of maxima and minima as the clearance is increased, but this theory fails at grazing angles and in the region of the nulls because the earth's surface is not plane. The diffraction loss over a knife-edge obstruction results in relatively minor oscillations around the free-space field as the clearance is increased, but shows a 6-db loss at grazing angles; this theory, in effect, assumes

unlimited antenna heights and a zero reflection coefficient.

On the other hand, the received field beyond the horizon of a perfectly smooth sphere decreases exponentially (linearly in decibels) and depends to some extent on the antenna heights. The dashed lines connecting these data on the figure with the curve for reflection from a plane surface show the region where neither of the three theoretical approaches is quite applicable and some interpolation is necessary. It will be noted that the smooth-sphere solution tends to approach the solution for knife-edge diffraction as the antenna heights increase.

The exponential decrease in field intensity beyond the line of sight in the smooth-sphere solution is caused by phase cancellation, and the net result is critically dependent on the smoothness of the sphere in much the same way as the results of the plane earth theory are critically dependent on the flatness of the plane. More exactly, the particular phase distribution that is produced by the smooth sphere in the region above the earth can be changed significantly either by the lack of a smooth surface or by the lack of a homogeneous atmosphere, but only the former effect is considered in this note. Variations in the terrain tend to upset this critical phase balance with the result that the field intensity is usually increased. Within the

line of sight a rough surface diffuses the reflected "ray," and thereby reduces the effective reflection coefficient and prevents deep cancellations; beyond the line of sight a rough surface has a similar effect, but it is more difficult to visualize because of the failure of the ray concept.

When the deviations from a perfectly smooth surface are very small compared with the first Fresnel zone, the smooth earth theory is expected to be applicable, but when the variations are comparable to or greater than the first Fresnel zone, the effective reflection coefficient approaches zero and, in the limit, the mean field intensity is expected to approach the value indicated by the knife edge or "rough sphere" diffraction theory.

Some experimental results in the 4,000- to 4,600-Mc range are shown on the figure. The circles were obtained on a very smooth path (Utah Salt Flats) by varying the antenna heights at both ends simultaneously. The crosses were measured on a "rough" path by varying only the antenna height at the New York end of the circuit. A profile of the latter path is shown in a recent paper by Durkee.<sup>1</sup> Mean values of ground wave

<sup>1</sup> A. L. Durkee, "Results of microwave propagation tests," Proc. I.R.E., vol. 36, pp. 197-205; February, 1948.

\* Received by the Institute, February 28, 1950.

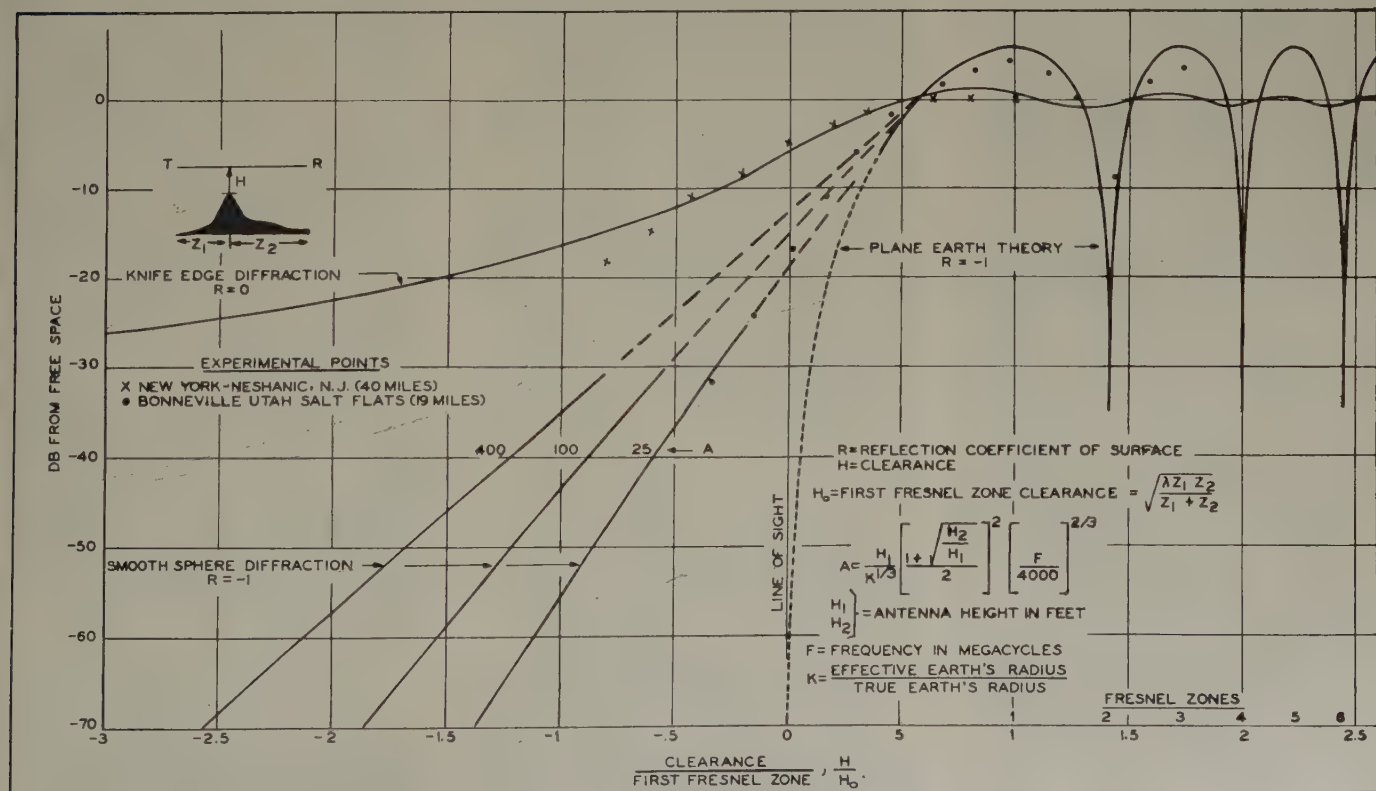


Fig. 1—Effect of clearance on radio transmission.

field intensity that are greatly in excess of those predicted by the smooth sphere theory have also been observed at much lower frequencies, notably in the 40- to 50-Mc range.<sup>2</sup>

This method of presenting radio propagation theory emphasizes both the importance of clearance and the great range of values of field intensity that can be obtained beyond the horizon. It presents an explanation of the enhancement of field intensity beyond the horizon to values greatly in excess of those calculated on the assumption that the surface of the earth is perfectly smooth. In addition, the rough sphere concept suggests that microwave transmission at distances far beyond the horizon may have sufficient strength and reliability to be useful for communication purposes, when reasonable increases in transmitter power become available. The Bell Telephone Laboratories plans to test this possibility.

KENNETH BULLINGTON  
Bell Telephone Labs.  
New York, N. Y.

<sup>2</sup> K. Bullington, "Radio propagation variations at vhf and uhf," *Proc. I.R.E.*, vol. 38, pp. 27-32; January, 1950.

### Suggestions to Technical Writers\*

Robert Hamlett's article on "Suggestions for the Preparation of Technical Papers" in the March, 1950, issue of the *PROCEEDINGS OF THE I.R.E.* is very well done.

There are, however, some minor facets of his presentation that might be better illuminated; for example, that paragraph which refers to personal pronouns.

It just so happens that the article in the February issue of *Physics Today* on "Physicists and the English Language," by William Fuller Brown, Jr., provides these very well-illuminated facets of this subject. Paper-writing engineers would do well to read it also, because it recommends a much less stuffy and pedantic style.

Engineers would make their papers much more readable and more easily digestible if they restrained their natural inclinations to exhibit their own profundity by omitting long-winded mathematical approaches where they can tell their stories in physical terms.

Who among us takes the time to check the involved mathematical presentations of most IRE papers, even if they be squarely in our own special field of interest? We are not carving out the ten commandments in indestructible stone tablets for the guidance of all eternity! We are merely trying to explain to our fellow engineers what we have found in some particular crevice of nature's secrets, so why not be a little more natural and direct when we set this down on paper? Besides, if our paper-publication object is to disclose and disseminate new information widely, we will never accomplish this by involved, abstract mathematics, where the language of physics will do the same thing better.

\* Received by the Institute, March 25, 1950.

If we are falling into the mathematical footsteps of the nuclear physicists, let us not forget that, beginning in the cob-webbed cubicles of the ancient alchemists, it has taken two thousand years to transmute the elements! Immanuel Velikovsky's Preface to his learned and provocative "Worlds in Collision" aptly says: "This book is written for the instructed and uninstructed alike. *No formula and no hieroglyphic will stand in the way of those who set out to read it.*" and "If . . . evidence does not square with formulated laws, it should be remembered that a law is but a deduction from experience or experiment, and therefore laws must conform with facts, not facts with laws."

BENJAMIN MIESSNER  
Miessner Inventions, Inc.  
Morristown, N. J.

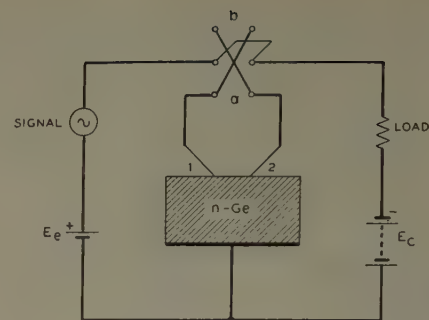


Fig. 1.—Circuit demonstrating reversible properties of an *n*-germanium transistor amplifier. When switch is in position *a*, electrode 1 is emitter, 2 is collector. When switch is in position *b*, 2 is emitter, 1 is collector.

Representative properties of such a transistor appear in Fig. 2. It may be seen that transmission is substantially the same in each direction at comparable values of emitter bias current.

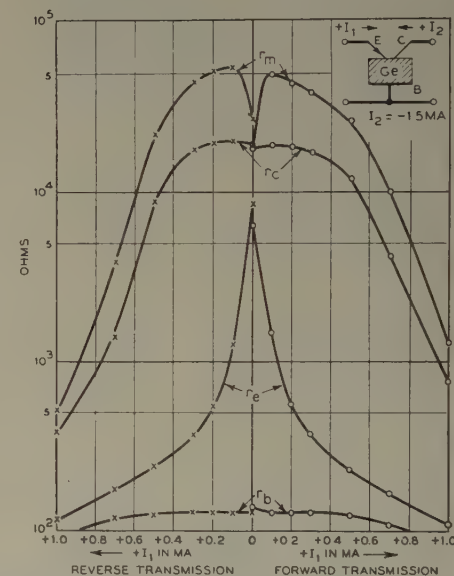


Fig. 2.—Small-signal impedances of a reversible *n*-germanium transistor. See footnote reference 4 for definitions of symbols.

In the *p*-germanium transistor the emitter is formed<sup>3,4</sup> and in this case also it is feasible to produce reversible transistor amplifiers.

The term "reversible" rather than "bilateral" has been used herein because at any given instant the device is unilateral, even though the amplifying direction can be reversed by changing the electrode biases.

W. G. PFANN  
Bell Telephone Laboratories, Inc.  
Murray Hill, N. J.

### The Transistor as a Reversible Amplifier\*

In the transistor described by Bardeen and Brattain two closely spaced electrodes of the point-contact type rest on the surface of a block of *n* germanium which has a large-area base electrode.<sup>1,2</sup> The point electrodes have been designated emitter and collector and, in conjunction with the base, they comprise the input and output terminals, respectively, of the transistor.

An interesting feature of the transistor, especially as compared with a triode vacuum tube, is the dual nature of the point electrodes. Either can be used as emitter or as collector, provided it is properly biased. For an electrode to function as a collector it is usually biased in the reverse, or high-impedance, direction, while to act as an emitter it is biased in the forward, or low-impedance, direction.

An electrical treatment known as "forming" has been described for the *n*-germanium transistor which increases the effectiveness of an electrode as a collector.<sup>2,3</sup> Such forming need not seriously impair the operation of an electrode as an emitter. Hence, it is feasible to form both rectifying junctions and to obtain thereby an improved, reversible transistor amplifier. By proper control of the forming operations, the properties of the device can be made substantially the same in both directions. A number of such transistors have been made in the laboratory.

The operation of a reversible *n*-germanium transistor is illustrated by the circuit of Fig. 1, which, however, is not intended to represent an actual application.

\* Received by the Institute, May 25, 1950.  
<sup>1</sup> J. Bardeen and W. H. Brattain, "The transistor, a semi-conductor triode," *Phys. Rev.*, vol. 74, p. 230; July 15, 1948.  
<sup>2</sup> J. Bardeen and W. H. Brattain, "Physical principles involved in transistor action," *Phys. Rev.*, vol. 75, pp. 1208-1225; April 15, 1949. Also *Bell Sys. Tech. Jour.*, vol. 28, pp. 239-277; April, 1949.  
<sup>3</sup> J. Bardeen and W. G. Pfann, "Electrical forming of *n*- and *p*-germanium transistors," *Phys. Rev.*, 1950.

<sup>4</sup> W. G. Pfann and J. H. Scaff, "The *p*-germanium transistor," *Proc. I.R.E.*, this issue, pp. 1151-1154. Also *Phys. Rev.*, vol. 76, p. 459; August, 1949.

# Institute News and Radio Notes

## TECHNICAL COMMITTEE AND PROFESSIONAL GROUP NOTES

The Standards Committee, under the Chairmanship of J. G. Brainerd, held a meeting on July 20. Arthur Van Dyck, IRE Representative on ASA Section Committee Z-17 (Preferred Numbers), will write a paper on Preferred Numbers for publication in the PROCEEDINGS OF THE I.R.E. A Subcommittee of the Standards Committee has been formed for the purpose of defining terms of a general nature not in the field of any one technical committee. The list of terms collected will be submitted to the Standards Committee and given to the various technical committees for defining. If, upon presentation of these terms to the Standards Committee, it is found that no particular committee wishes to define them, they will then be defined by this Subcommittee. The Standards Committee approved the following standards: "Proposed Standards on Wave Propagation, Definitions of Terms, 1950," and "Standards on Television, Methods of Measurement of Time of Rise, Pulse Width and Pulse Timing of Video Pulses in Television." Both of these standards are scheduled for publication in the November issue of the PROCEEDINGS OF THE I.R.E. A supplement to the "Standard on Television: Methods of Measurement of Television Receivers, 1948," was also approved. This supplement covers intercarrier sound receivers. It will not be published separately but a mimeographed copy will be included with each copy of the 1948 Standards. "Definitions in Network Topology," prepared by the Circuits Committee, received approval of the Standards Committee. The publication date for these definitions has not been scheduled. The IRE will participate in a Joint IRE/AIEE/RDB Conference on Electron Tubes for Computers which is expected to be held in the Fall. Complete details will be announced. A new committee, to be known as the Joint IRS Committee for Television, has been formed by IRE, RTMA, and SMPTE, to avoid duplication of effort in the preparation of standards in the fields associated with television. The initials "IRS" represent the first letter of each sponsor's title. Axel G. Jensen, Millard W. Baldwin, Jr., and R. L. Garman were appointed as the IRE Representatives to the Committee. . . A Joint IRE/AIEE Committee on Noise Definitions has also been organized with the specific task of critically reviewing all existing noise definitions and adding any such new definitions as might be required. The IRE personnel on this Committee is as follows: Stanford Goldman, Syracuse University; Jerry Minter, Measurements Corporation; A. W. Friend, RCA; Ray Moffett, NBC; Claude Shannon, BTL; and R. M. Ryder, BTL. . . The Third Annual Joint IRE/AIEE Conference on Electronic Instrumentation in Nucleonics and Medicine will be held on October 23, 24, and 25, at the Hotel Park Sheraton, New York, N. Y. For details on the Conference, see page 1224. . . The re-

cently organized IRE Professional Group on Radio Telemetry will now be known as the IRE Professional Group on Radio Telemetry and Remote Control. . . The IRE Professional Group on Vehicular and Railroad Radio Communications has also changed its title to the IRE Professional Group on Vehicular Communications. . . The Joint Technical Advisory Committee held a meeting on August 3 under the Chairmanship of J. V. L. Hogan. Table #1, which appears in Volume 4 of JTAC, has been revised.

### NOTICE

Effective January 1, 1951, the Student dues will be \$5.00 per year. Payments at the old rate of \$3.00 will be accepted through December 31, 1950. All payments received thereafter will be at the new rate.

### Signal Corps Center Recruits Civilian Technical Personnel

The U. S. Signal Corps Center Headquarters at Fort Monmouth, N. J., has announced position vacancies for communication and electronic engineers, teachers, and technical writers in grades ranging from GS-5 to GS-12. Civilian technical personnel with extensive professional experience who can qualify for the higher grades in these categories are particularly needed.

For further information and for making application, write to the Civilian Personnel Branch, Building T-530, Fort Monmouth, N. J.

Applicants should submit a completed Standard Form 57, "Application for Federal Employment," for review before coming to Fort Monmouth for a personal interview. Standard Form 57 can be obtained at a first or second class post office.

### 3-DAY PROGRAM ANNOUNCED FOR RADIO FALL MEETING

The three-day Radio Fall Meeting, sponsored jointly by the RTMA and IRE, will convene on Monday, October 30, at the Hotel Syracuse, Syracuse, N. Y. The tentative program, announced by Virgil M. Graham, chairman, is as follows:

Monday, October 30, 1950

9:30 A.M.—General Session, R. R. Batchner, Presiding. Welcome by Chairman of Syracuse IRE Section; "A Broad Range Oscillator for Television Testing," by H. A. Finke and J. Ebert, Polytechnic

Research and Development Corp.; "Wide-Band Impedance Matching Between a Resonant Antenna and a Line," by H. A. Wheeler, Wheeler Laboratories, Inc.; "High-Frequency Crystal Calibrator Design," by J. B. Minter, Measurements Corporation; "The Determination of Amplifier Sensitivity With the Aid of the Noise Diode," by W. K. Squires, Sylvania Electric Products Inc.

2:00 P.M.—Quality Control Session, J. R. Steen, Presiding. (Sponsored by IRE Professional Group on Quality Control.) "Application of Statistical Quality Control in Manufacture of Electronic Products," by E. R. Ott, Rutgers University, and George Scheel, Sonotone Corp.; "The Quality Control Indicator," by C. J. Falk, General Electric Co.; "The Control of Averages in Radio Tube Manufacture," by A. K. Wright, Tung-Sol Lamp Works, Inc.; "The Human Aspect of Engineering Quality into the Product," by Carl Gartner, Allen B. DuMont Labs., Inc.

8:00 P.M.—Joint Session with Technology Club of Syracuse.

Tuesday, October 31, 1950

9:00 A.M.—Television Session, R. A. Hackbusch, Presiding. (Sponsored by the IRE Professional Group on Broadcast and Television Receivers). "A Study of Permanent Magnet Focusing Devices for Television Picture Tubes," by Kenneth James and R. T. Capodanno, Emerson Radio & Phonograph Corp.; "The Application of a New Low-Noise Double Triode as an RF and IF Amplifier in Television Receivers," by R. M. Cohen, RCA; "Consideration of Optimum Use of Picture Tubes," by W. B. Whalley, Sylvania Electric Products Inc.; "Evaluation of Performance Characteristics of Cathode-Ray Tubes for Use in Television Receivers," by K. A. Hoagland, Allen B. DuMont Labs., Inc.

2:00 P.M.—Television Session, D. D. Israel, Presiding. (Sponsored by IRE Professional Group on Broadcast and Television Receivers). "The Technical Aspects of Phonevision," by E. M. Roschke, Zenith Radio Corp.; "An Analysis of Color Television," by A. V. Loughren, Hazeltine Electronics Corp.

6:45 P.M.—Radio Fall Meeting Dinner. Toastmaster, W. R. G. Baker; Speaker, R. C. Sprague, President of RTMA.

Wednesday, November 1, 1950

9:00 A.M.—Audio Session, O. L. Angevine, Jr., Presiding. (Sponsored by the IRE Professional Group on Audio). "The Mechanics of the Phonograph Pickup," by T. E. Lynch, Brush Development Co.; "Lightweight Pickup and Tone Arm," by C. R. Johnson and L. J. Anderson, RCA; "Sound Pickup in High Ambient Noise," by Wayne Beaverson, Electro-Voice, Inc.; "RTMA Standards For Sound Equipment," O. L. Angevine.

# Institute News and Radio Notes

## NUCLEAR-MEDICAL INSTRUMENT CONFERENCE PROGRAM ANNOUNCED

Over 800 persons are expected to attend the Third Annual Joint AIEE/IRE Conference on Electronic Instrumentation in Nucleonics and Medicine to be held on October 23, 24, and 25, 1950. The location of this meeting, which will feature technical papers and discussions on current problems and advances in the vital fields of nuclear and medical science, is the Park Sheraton Hotel, New York, N. Y.

During the last two days of the conference, many interesting and informative exhibits of instruments and related products will be displayed by leading companies in the field. Those attending the meeting will have an opportunity to see many of the devices, some in actual operation, which will be discussed in the technical papers.

Highlighting the conference will be a special evening round-table discussion during which a number of prominent authorities will discuss "The Effects of Atomic Weapons," a government publication containing recently declassified information on this vital and timely subject.

The registration fee for the conference is \$3.50 per person. All papers and discussions presented during the meeting will be published in a proceedings of the Conference, available at a later date for \$4.00 per copy.

The tentative program of the Conference has been announced as follows:

### Monday, October 23, 1950

- 10 A.M.—"The Needs of Physiology and Medicine for Better Instrumentation for the Measurement of Respiratory Gases," O. Fenn, University of Rochester, Rochester, N. Y.; "Analysis of Respiratory Gases with Mass Spectrometer," A. Hitchcock, Ohio State University, Columbus, Ohio; "The Application of the Infrared Spectrophotometer to the Analysis of Respiratory Gases," E. D. Palmes, New York University, Bellevue Medical Center, New York, N. Y.; "The Measurement of Oxygen in Gases by Paramagnetism," O. Beckman, South Pasadena, Calif.
- 2 P.M.—"The Delineation of Intracranial Structure with the Aid of Ultrasonic Waves," R. H. Bolt, H. T. Ballantine, G. R. Ludwig, and T. E. Hueter, MIT, Cambridge, Mass.; "The Localization of Brain Tumors with Radioactive Isotopes," Theodore Fields, Veterans' Hospital, Hines, Ill.; "The Use of Isotopes in the Measurement of Body Fluids," J. L. Nickerson, Columbia University, New York, N. Y.; "Developing of Sectioning Techniques for Electron Microscopy," (to be read by S. G. Ellis), James Hillier, RCA Laboratories, Princeton, N. J.

### Tuesday, October 24, 1950

- 10 A.M.—"General Survey of Health Physics Instrumentation Problems," H. M. Parker, General Electric Co., Hanford Works, Richland, Wash. "Calibration of Radiation Detection Instruments," L.

D. Marinelli, Argonne National Laboratory, Chicago, Ill.; "A Calorimetric Method of Measuring High X-Ray Intensities," W. T. Ham, Jr., Medical College of Virginia, Richmond, Va.; "Consideration of Radiation from High-Voltage Cathode-Ray Tubes," O. W. Pike, General Electric Co., Schenectady, N. Y.

- 2 P.M.—"Fast Neutron Dosimetry and Related Problems," G. S. Hurst and R. H. Ritchie, Oak Ridge National Laboratory, Oak Ridge, Tenn.; "New Medical Applications of Tracers," A. H. Holland, Armour Research Foundation, Chicago, Ill.; "Electronics and Nucleonics Applied to Enzymes and Viruses," E. C. Pollard, Yale University, New Haven, Conn.

Following the afternoon session will be the Annual meeting of the IRE Professional Group on Nuclear Science.

- 8 P.M.—"Round Table Discussion on 'The Effects of Atomic Weapons'." Speakers to include Brig. Gen. J. P. Cooney, AEC; H. D. Bowman, Drexel Institute and Consultant to AEC; and Herbert Scoville, Armed Forces Special Weapons Project.

### Wednesday, October 25, 1950

- 10 A.M.—"Manufacture and Quality Control of Geiger Tubes," D. L. Collins, Victoreen Instrument Co., Cleveland, Ohio, and D. Atchley, Tracerlab, Boston, Mass.; "Boron-Lined Neutron Proportional Counters," W. W. Schultz, General Electric Co., Schenectady, N. Y.; "Design of a Commercial Scintillation Counter," E. W. Jarvis, Jr., W. S. MacDonald Co., Cambridge, Mass.; "Testing Photomultipliers for Scintillation Counting," R. W. Engstrom, RCA Laboratories, Princeton, N. J.
- 2 P.M.—"Scintillation Counter Instrumentation," G. Cowper, Chalk River, Ont., Canada; "Recent Advances in Electron Techniques in Canada," N. F. Moody, Chalk River, Ont., Canada; "Fast Counting," Martin Deutsch, MIT, Cambridge, Mass.; "New Developments in Mass Spectrometry," John Hipple, National Bureau of Standards, Washington, D. C.

## GENERAL ELECTRIC COMPANY DEVICES NEW TYPE COUNTER

A new instrument for the detection and counting of alpha, beta, and gamma particles has been announced by General Electric's special products division. Called a Universal scintillation counter, the device provides low background alpha counting and high-efficiency beta and gamma counting, according to engineers of the company's General Engineering and Consulting Laboratory who worked on its development.

It is intended for use in counting samples and smears in health physics work, for analytical determination of disintegration rate, and in counting radiation of ore samples. The sampling chamber can accommodate specimens up to two inches in diameter.

## Calendar of

## COMING EVENTS

- National Academy of Sciences Meeting, G. E. Research Laboratory, Schenectady, N. Y., October 9-11
- SMPTe Semiannual Convention, Lake Placid, N. Y., October 16-20
- IRE-AIEE Conference on Electronic Instrumentation in Nucleonics and Medicine, Hotel Sheraton, New York, N. Y., October 23-25
- Audio Fair, Sponsored by Audio Society of America, Hotel New Yorker, New York, N. Y., October 26-28
- Optical Society of America Meeting, Hotel Statler, Cleveland, Ohio, October 26-28
- Radio Fall Meeting, Syracuse, N. Y., October 30, 31, November 1
- National Conference of the IRE Professional Group on Vehicular Communications, Detroit, Mich., November 3
- IRE-AIEE Conference on Electron Tubes for Computers, Washington, D. C., December 14-15
- AAAS Annual Meeting, Cleveland, Ohio, December 26-30
- 1951 IRE National Convention, Waldorf-Astoria Hotel, New York, N. Y., March 19-22
- IRE Southwestern Conference, Dallas, Texas, April 20-21, 1951

## SECOND CALL!

## AUTHORS FOR IRE NATIONAL CON- VENTION!

E. Weber, Chairman of the Technical Program Committee for the 1951 IRE National Convention, requests that prospective authors submit the following information:

1. Name and address of author
2. Title of paper
3. A 100-word abstract and additional information up to 500 words (both in triplicate) to permit an accurate evaluation of the paper for inclusion in the Technical Program.

Please address all material to E. Weber, Microwave Research Institute, Polytechnic Institute of Brooklyn, 55 Johnson Street, Brooklyn 1, N. Y. The deadline for acceptance is November 20, 1950. Your prompt submissions will be appreciated.

# IRE People

**Frederick A. Kolster** (A'12-M'13-F'16-L'49), who had been engaged as a consulting engineer at



F. A. KOLSTER

San Francisco, Calif., of later years, and who was an early and prominent member of the Institute, passed away in July. A pioneer specialist in the fields of radio engineering and radio physics, Mr. Kolster earned recognition for his contributions to the development of radio.

Born on January 13, 1883, at Geneva, Switzerland, Mr. Kolster was educated in New England public schools and was graduated from Harvard University in 1908.

He was an assistant to John Stone Stone and Lee de Forest in the early days of radio, and an assistant to Fritz Lowenstein in 1911 and 1912. Until 1921 Mr. Kolster was chief of the radio section of the National Bureau of Standards, Washington, D. C., and later served as Consultant at the Bureau of Ships, Electronics Division, with the Navy Department in Washington.

Joining the Federal Telegraph Company in Palo Alto, Calif., in 1921 as a research engineer, he remained there until 1931 when he went to the International Telephone and Telegraph Corporation. He also was affiliated with the Radio Navigational Instrument Corporation.

Mr. Kolster was responsible for numerous radio devices, ranging from precision-measuring instruments to special forms of television antennas.

He was a Director of the Institute during 1933 and 1935, and was active on the following IRE Committees: Admissions, 1935-1936; Membership, 1930-1932; Nominations, 1933; Papers, 1913; Sections, 1933; Standardization, 1914-1921, 1925, 1927-1931; Subcommittee on Marine and Direction-Finding Receivers, Chairman, 1930-1931; Wavelength Regulation, 1916; and Wave Propagation, 1937-1938.

**Eric Walker** (M'47) became executive secretary of the Research and Development Board on August 1, and is on leave from Pennsylvania State College, where he is director of the Ordnance Research Laboratory and the head of the electrical engineering department.

Dr. Walker has been associated with the board during the past year as a consultant to its committee on electronics. He recently took a leave from the college to serve as deputy chairman of a weapons panel in the National Research Council. He has been with Pennsylvania State College since July, 1945, and during that time he inaugurated a series of annual conferences on the administration of research. These were attended by administrators of university and industrial research organizations.

During the war Dr. Walker was assistant director of the Underwater Sound Laboratory at Harvard University and served with the Office of Scientific Research and Development. From 1940 to 1942 he was head of the Electrical Engineering Department of the University of Connecticut and served in the same capacity at Tufts College, Medford, Mass., from 1934 to 1940.

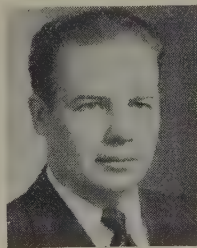
While with the University of Connecticut he was in charge of research for the electrical division of the Colt's Patent Firearms Manufacturing Company, and while at Tufts College was consulting engineer for the Doble Engineering Company as well as technical assistant to the president.

Dr. Walker serves as consultant to the Koppers Company, Pullman-Standard Car Manufacturing Company, and the Mine Safety Appliances Company, and is a licensed professional engineer in the States of Pennsylvania and Connecticut. He has invented a polar vector indicator and a gallstone detector, and is the author of several books and papers in technical magazines.

He was born in Long Eaton, England. Dr. Walker holds the Doctor of Science degree from Harvard University. He is a Fellow in the AIEE, the Acoustical Society of America, and the American Physical Society; and is a member of the American Society for Engineering Education and the American Institute of Physics. In 1948 he was awarded the President's Certificate of Merit for outstanding service for the Armed Forces during World War II.



**Donald E. Smith** (A'37), formerly an engineer for the electronics division of Sylvania Electric Products Inc., Boston, Mass.,



DONALD E. SMITH

has been transferred to the renewal tube sales department of the radio tube division. He will serve as a sales representative operating out of Emporium, Pa., but will be located at St. Louis, Mo.

Mr. Smith joined the Sylvania engineering staff at the Product Development

He received the B.S. degree in electrical engineering from Purdue University, where he was graduated with distinction. Mr. Smith is a member of Tau Beta Pi and Eta Kappa Nu.

**John F. Bates** (A'36-SM'45), instructor in physics at the University of Illinois, died recently. He was a native of Philadelphia, Pa., and was born on November 24, 1892.

He was graduated from the U. S. Naval Academy in 1915 with the B.S. degree, and attended Harvard University. From 1930 through 1932 he was an instructor in radio engineering, crypt analysis, and communications, at the postgraduate school at Annapolis, and for the following two years was on service at the Underwater Sound Research on the *U.S.S. Eagle*. From 1935 to 1936 he was electronics assistant to the director of the Naval Research Laboratory at Anacostia, Washington, D. C. He then was an inspector of electronics materials at RCA in Camden, N. J., and electronics director of all radio contractors in the Chicago area.

**Ralph A. Hackbusch** (A'26-M'30-F'37), president and managing director of Stromberg-Carlson Co. Ltd., Canada, was elected



R. A. HACKBUSCH

president of the Radio Manufacturers Association of Canada at the twenty-first annual meeting held in June. The Manufacturers also elected A. B. Hunt (A'43-SM'43), manager of communications equipment division of the Northern Electric Company Ltd., Montreal, as vice-president of the association, and re-appointed S. D. Brownlee as executive secretary.

Mr. Hackbusch has been associated with development of the radio industry in Canada for many years, and was elected vice-president of the Radio Manufacturers Association two years ago.

From October, 1940, to 1943, Mr. Hackbusch was vice-president and director of the Radio Division of the Canadian Government Research Enterprises Ltd. Mr. Hackbusch returned to Stromberg-Carlson Company Ltd., as vice-president and managing director in November, 1943, and was elected president in 1949. Mr. Hackbusch is also president of the Canadian R.T.P.B., a member of the Radio Club of America, and a member of the Association of Professional Engineers of Ontario. He was vice-president of the IRE in 1944.

**Nicholas M. Oboukhoff** (M'23-SM'43), research professor emeritus of electrical engineering and professor emeritus of mathematical physics at Oklahoma Agricultural and Mechanical College, died recently at Stillwater, Okla. He was 77.

Mr. Oboukhoff received the physics and mathematics degree from the University of Moscow, Russia, and the degree of engineer technologist, from the Technological Institute of Kharkov, Russia. He earned the Ingenieur Electricien Diplome E.S.E., Ecole Supérieure d'Electricité de Paris, France, and the Ph.D. cum laude from the California Institute of Technology. He was a special student in philosophy at the University of Chicago.

He was a designer with the Locomotive Building Works of Kharkov, Russia, during 1904 and 1906, and then went on a technical inspection and study tour through France, Switzerland, England, Belgium, Italy, Austria, Germany, and Egypt. He inspected irrigation and Nile navigation control, as well as irrigation problems in Palestine.

During the years 1911 through 1941, he taught in a trade school in Paris and then became a research engineer. He was associated with several concerns in Petrograd (now Leningrad), among them Siemens and Halske Co. He was Commissioner of Labor in Donetz District and Ural District, Russia, under the Provisional Government during 1918 and 1919. Mr. Oboukhoff was then put in charge of the Irkut River Project (the hydroelectric development of the river). He held patents in France, the United States, and England, dealing chiefly with medium and high-frequency alternators. He was also the author of numerous technical papers published in different languages.

Mr. Oboukhoff was a member of the Sigma Xi and Eta Kappa Nu, and a Fellow of the American Association for the Advancement of Science, the American Geographical Society and the Oklahoma Academy of Science. He was also an accomplished musician.



**Homer M. Sarasohn** (A'41-M'48) has returned to the United States after spending more than three years at General MacArthur's headquarters in Japan. During his stay, he has assisted and advised the Supreme Commander on communication matters through the Civil Communications Section, and has also assisted the Japanese through that section, as well as helping them in the solution of their industrial engineering problems. His special mission has been rehabilitation and reorientation of the Japanese

communications equipment manufacturing industry.

At the time of Sarasohn's arrival with the Occupation Force, he found that 85 per cent of this industry had been destroyed during the war. This, together with the relatively inadequate level of engineering in Japan at that time, and the feudalistic management practices being used, was a severe obstacle to the re-establishment of a completely stable nation.

He directed his efforts towards the reorganization of the industry and introduced the modern concepts and practices of statistical quality control. His activity was also related to the Japanese Government. He assisted the Ministry of Telecommunications, which owns and operates the telephone and telegraph systems of Japan, the Broadcasting Corporation of Japan, which is government-supported, and the National Rural Police in the revision and modernization of their communications equipment, and also of their procurement specifications and procedures.

Formerly, Mr. Sarasohn was engaged by the Raytheon Manufacturing Company, Waltham, Mass., in its microwave research activity.



**E. H. Ulm** (M'46), formerly sales engineer for the electronics division, Sylvania Electric Products Inc., has been appointed merchandising manager of the company. He joined the staff of the electronics division of Sylvania Electric in 1945 as a sales engineer.



E. H. ULM

Prior to that time he was associated with the field engineering force of radio division of the Western Electric Company, where he served as an instructor in radar and sonar. During 1943 he served as an antisubmarine warfare field engineer for the Division of War Research, Columbia University.

Mr. Ulm, who is a native of Fort Dodge, Iowa, was graduated from Carleton College in 1938 and later did postgraduate work at Carleton College, the University of Iowa, and Northwestern University. He is also a member of the AIEE and the Radio Club of America.



**Robert E. Martin** (S'49), who received the master of science degree and the bachelor of electrical engineering degree from Ohio State University at the recent Spring Commencement, has been appointed to the staff of Batelle Institute, Columbus, Ohio. He will be engaged in research in electrical engineering.



**James Winston Woody, Jr.** (S'49) who recently received the E.E. degree from the University of Kentucky, has joined the staff of the instrument department, Engineering Division, of the Oak Ridge National Laboratory, Oak Ridge, Tenn.

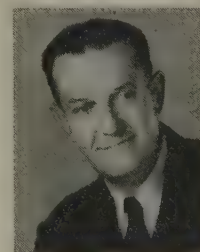
**Norman R. Beers** (SM'50), who formerly had been in charge of the meteorology group at the Brookhaven National Laboratory, Upton, L. I., N. Y., and was the editor of *Nucleonics*, a McGraw-Hill publication, until his resignation owing to ill health, died recently at Fort Hamilton Veterans' Hospital. Mr. Beers was 38 years old.

A native of Missouri, he attended the University of Missouri, and from 1934 to 1937 he was a Rhodes Scholar at Oxford University, specializing in astrophysics. He was associated with the college textbook department of McGraw-Hill until 1941, when he joined the Navy. He was on active duty for five years. Mr. Beers was in charge of the basic-phase Aviation Machinist Mate School at the Naval Air Station, Jacksonville, Fla., and was later in the Bureau of Aeronautics, in Washington, and on the faculty of the postgraduate school of the United States Naval Academy, at Annapolis, Md. He was a lieutenant commander at the time of his discharge from the service.

He remained on the Annapolis staff as associate professor of aerological engineering until 1947, when he joined the Brookhaven atomic project.

In 1949 he became editor of *Nucleonics* and continued as a consultant for the Brookhaven laboratory. Mr. Beers was co-author of the McGraw-Hill "Handbook of Meteorology," and had published a number of papers in technical journals.

**Rawson Bennett** (A'36-M'43-SM'43-F'50), who has been commanding officer and director of the U. S. Navy Electronics Laboratory, San Diego, Calif., for four years, reported in August to the Armed Services Industrial College in Washington, D. C. Captain Bennett was named director of the Laboratory in 1946.



RAWSON BENNETT

Prior to his assignment as director of the Navy's electronics research and development installation on Point Loma, Captain Bennett was head of the design branch, electronics division, of the Bureau of Ships. He holds the Navy Department's Legion of Merit Award for his contributions to varied developments in the military field of submarine detection.

Graduated from the Naval Academy in 1927, Captain Bennett received the Master of Science degree in engineering from the University of California at Berkeley in 1937. In March, 1950, he was elected a Fellow of the Institute, becoming the fifth officer of the regular armed services to receive that honor in the history of the IRE.

# Industrial Engineering Notes<sup>1</sup>

## TELEVISION NEWS

Sales of cathode-ray tubes to television receiver manufacturers in May increased 20 per cent over the preceding month as the trend to larger screens in home receivers was further emphasized, according to RTMA member-company reports. More than 64 per cent of all TV receiver-type tubes sold to manufacturers during the month were 14 inches and over in size. The previous dominant tube size—12 through 13.9 inches—represented only 35 per cent of the total picture tubes sold to manufacturers in May. . . . The Hazeltine Electronics Corp., in conclusions on color television offered to the FCC, declared that none of the proposed color television systems "currently is really ready" for standardization. The FCC did not formally accept the company's conclusions as part of the color record, as Hazeltine was not a party to the hearing and its proposals did not conform to FCC procedure. Hazeltine asked the FCC to take two steps in arriving at its color decision. First, the company suggested the FCC make a broad policy decision and then follow this action with close co-operation with industry in order to get detailed standards. . . . Television receiver production during the first half of 1950 equalled that of the entire year of 1949, attaining a new industry record of almost 3,100,000 sets up to July 1, according to RTMA estimates based on reports by member companies. Radio receiver production also was up, about 50 per cent higher than during the corresponding six-month period last year. . . . The General Electric Company has advised the FCC that it had developed a new color television system which has "very important advantages both technically and economically." W. R. G. Baker, vice-president, wrote Chairman Wayne Coy that the new system, identified as frequency interlace, "appears to have such outstanding potential advantages as to justify serious consideration even at such a late date." In a detailed description of the new system, General Electric said: "Among the apparent technical advantages are compatibility with existing monochrome standards and freedom from twinkle, crawl, flicker, color fringing, and transient color shifts."

## MOBILIZATION

Military procurement of electronic equipment and components during the current fiscal year are expected by key defense officials to exceed \$1 billion and may run as high as \$1.5 billion or more, according to information obtained by RTMA. While an accurate breakdown of military requirements

from the radio-television industry is not yet available, officials point out the following: Regular military appropriations measure of more than \$13 billion carries funds for electronic procurement of between \$500 and \$600 million; electronic procurement in the special \$10 billion budgetary request will add another \$500 million to \$1 billion; military orders for electronic items already let under the past fiscal year's appropriations add another \$300 to \$400 million; and an additional appropriation for further rearmament of American allies, as forecast by the President, unquestionably will include additional funds for electronic procurement. This over-all appropriation, according to some estimates, may run as high as \$10 billion. . . . While the FCC was continuing its day-to-day deliberations preparatory to issuing a decision in the color TV proceedings, there was speculation in Washington as to what effect the international situation may have on the Commission's findings both in this case and the forthcoming uhf allocation inquiry and the lifting of the TV "freeze." Observers point out that the FCC's color decision may become "academic" or have to be shelved until the present rearmament program is completed, in view of the anticipated military demands on the production capacity and engineering talent of the radio-television industry. There appeared to be general anticipation in Government circles that production of TV receivers will be sharply cut back within the next few months by the mere impact of military procurement and industrial controls on vital materials and priorities, regardless of whether TV and radio receiver production is reduced by actual limitation orders. Curbs on consumer credit are expected to slow down TV set buying concurrent with reduced production. It is also pointed out that Administration policies designed to halt "expansion" of consumer industries may result in a practical continuation on the "freeze" on new TV station construction regardless of the action of the FCC. On the other hand, however, some officials believe that expansion of TV reception, even in wartime, would be desirable from the Government's standpoint because of the value of the medium in transmitting important information to the public.

## SECRET GENERAL MANAGER OF RTMA AS GEDDES RETIRES AFTER 23 YEARS

James D. Secrest, who for several years has been Director of Public Relations for the Radio-Television Manufacturers Association, assumed the duties of secretary and general manager of RTMA on August 1, succeeding Bond Geddes, who retired as executive vice-president after 23 years of service with the association. Mr. Geddes continues to serve as a consultant under an arrangement with the Board of Directors.

A long-time newspaper man in Washington, D. C., Mr. Secrest was associated with newspapers in Cincinnati, Ohio, and Asheville, N. C., before coming to Washington in 1929. From then until 1941 he was on the staff of the *Washington Post*, during which time he was also a member of the Capitol staff. He also reported radio industry news.

Early in 1941 he joined the Information Division of the Office of Emergency Management which subsequently became the Office of War Information. He helped to organize and directed the OWI domestic field service comprising sixty offices throughout the United States.

Before joining RTMA in March, 1945, Mr. Secrest was in charge of publicity and advertising for the wartime pulpwood production campaign conducted by the War Activities Committee of the Pulpwood Consuming Industries, with headquarters in New York, N. Y.

## NBS DEVELOPS NEW OSCILLATOR

The National Bureau of Standards has announced the development of a resistance-capacitance oscillator "with many immediate applications in radio and electrical work." The new oscillator covers, in five steps, the frequency range from 20 cycles to 2 megacycles. In the new RC oscillator, a single amplifier driving a cathode follower, provides wide-band operation with small phase shift, low output impedance, and constant output voltage. Complete details of the new development are to be published in "The Technical News Bulletin," a monthly publication of the National Bureau of Standards.

## KAAR WILL REPRESENT RTMA AT SESSIONS OF CCIR STUDY GROUP

I. J. Kaar of General Electric Company, who is chairman of the Committee on Television, Receiver Section, RTMA Engineering Department, represented RTMA and the industry at a meeting of the CCIR Study Group 11 in Geneva, Switzerland, at the designation of Director W. R. G. Baker. He was appointed upon the recommendation of Donald S. Parris of the U. S. Department of Commerce, who was one of the American delegates to the CCIR sessions held recently in London, at which Donald G. Fink represented RTMA.

The RTMA Export Committee, at its recent meeting in Chicago, adopted a resolution recommending to the Board of Directors that RTMA be represented at all pertinent international meetings on radio and TV problems. The Geneva session is scheduled to give further consideration to FM versus AM for TV sound, positive or negative modulation, and other standards not agreed on at the London CCIR meeting.

## NEW NAME OF RTMA IS OFFICIAL; INSIGNIA FOLLOWS RMA DESIGN

The official name of the former Radio Manufacturers Association has now become the Radio-Television Manufacturers Association. Judge John W. Van Allen, RTMA General Counsel, has advised RTMA headquarters that the articles of amendment to the RTMA charter, as approved by the membership at the June meeting, have been filed with the State of Illinois and become effective immediately.

<sup>1</sup> The data on which these NOTES are based were selected by permission from *Industry Reports*, issues of July 7, July 14, July 21, and July 28, published by the Radio-Television Manufacturers Association, whose helpful attitude is gladly acknowledged.

# ATOMIC ENERGY COMMISSION WILL PUBLISH REPORTS TWICE MONTHLY

The Atomic Energy Commission and the Department of Commerce recently jointly announced that business firms will have easier access to nonsecret atomic energy technical reports under a new distribution plan. Under the plan, the Office of Technical Services of the Department of Commerce will become the sales agency and reference source for nonsecret AEC technical reports.

"*Nuclear Science Abstracts*," a magazine issued twice monthly, is the principal medium through which business firms may learn about reports relating to atomic energy research. The magazine, which provides titles and abstracts of publicly and privately published reports in these fields, sells for \$6 per year. Subscriptions should be addressed to the Office of Technical Services, U. S. Department of Commerce, Washington 25, D. C.

# SKIATRON CORPORATION DEVELOPS "SUBSCRIBER VISION" TV SYSTEM

The Skiatron Corporation of New York, N.Y., has informed the Federal Communications Commission that it expects to demonstrate this fall a "Subscriber Vision" television system by which paying subscribers would be enabled to view movies on their receivers.

"We are proceeding energetically on a pilot model," the New York concern wrote the FCC, "and following its completion, we intend to apply to the FCC for the same type of experimental public test privilege in New York City as the Zenith Radio Corporation has obtained in Chicago."

The firm claims to own the only "practical, patented system for providing a special television service to paying subscribers." Skiatron says its method provides for the broadcasting pictures over the air "without the necessity for any intervening telephone

connection." It said further that Skiatron intends to offer other manufacturers in the television industry licensing arrangements under its proposed system.

# J. V. L. HOGAN IS NAMED CHAIRMAN OF JTAC TO SUCCEED DONALD FINK

John V. L. Hogan has succeeded Donald G. Fink as Chairman of the Joint Technical Advisory Committee of RTMA and IRE. The group was organized by RTMA and IRE to advise the government on technical matters as a successor to the former Radio Technical Planning Board.

I. J. Kaar succeeds Mr. Hogan as vice-chairman of the committee. Other JTAC members are: Donald G. Fink, Ralph Brown, T. T. Goldsmith, Jr., Haraden Pratt, Philip F. Siling, and David B. Smith.

## Books

### The World's Radio Tubes (Brans, Vademecum), 1950 International Edition

Published (1950) by P. H. Brans, Ltd., Antwerp. Distributed by Editors and Engineers, Ltd., 1300 Kenwood Road, Santa Barbara, Calif. 508 pages +xxii +xx pages. 8 X 11½.

This is the eighth edition of the late P. H. Brans' handbook giving the principal rated characteristics of electron tubes manufactured throughout the world. Introductions, abbreviations, and instructions for using the book are given in most of the common European languages except Russian.

Tubes are listed by type number first, together with manufacturer's name, cathode heater or filament voltage, basing index and reference to a second table which gives rated electrode voltages, currents, and dynamic characteristics. Listed also are the type numbers of tubes having equivalent characteristics to facilitate substitution when the original tube cannot be duplicated; base connections are also given for the equivalent types.

Coverage, at least so far as American tubes are concerned, is rather complete and includes transmitting, cathode-ray, and special-purpose tubes, in addition to receiving tubes. In most cases a single manufacturer is listed, and one suspects that the editors have overlooked the fact that most American tubes having a type number assigned by RMA are made by several manufacturers and are listed in the catalogues of most of the larger concerns. This handbook will probably be of interest to persons in the United States who are interested in foreign tube types and in exporting. It is easily understandable that it would be especially useful in countries where most of the tubes used are imported.

GEORGE D. O'NEILL  
Sylvania Electric Products Inc.  
P.O. Box 6  
Bayside, L. I., N. Y.

### New Publications

**The Table of Complex Numbers** by Herbert E. Salzer, National Bureau of Standards Applied Mathematics Series AMS8, 44 large pages, is available from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C., for 25 cents a copy. Remittances from foreign countries should be in United States exchange and should include an additional sum of one-third the publication price to cover the cost of mailing.

The booklet has been published to meet the needs of engineers, mathematicians, and scientists who are working on problems in fields as diverse as alternating-current circuits, number theory, and exterior ballistics. The Table is especially useful in the treatment of Taylor series involving complex variables. It is the first extensive table of powers of complex numbers ever published.

In this Table, exact values of powers of complex numbers are given in Cartesian form for powers from 1 through 25 and for arguments with real and imaginary parts ranging separately from 0 to 10 in unit steps. The Table is arranged essentially in order of magnitude of the distances of the argument values from the origin in the complex plane.

**Colorimetry**, Circular C478, by Deane B. Judd, is a publication issued by the National Bureau of Standards and available from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C., at 30 cents a copy. The circular contains 56 large double-column pages and is illustrated.

Scientists and technicians concerned with color measurement will find much helpful information. For the past 20 years, the National Bureau of Standards has played a

leading part in establishing practical working standards on color and in setting up and administering color tolerances. As a result, conformity to a color requirement can now be determined with the same assurance as mechanical properties.

The booklet describes the standards and measurements methods developed by the Bureau in this field, giving the basis for each technique and showing clearly how one method supplements the other. Chapters cover the following topics: the standard observer; illuminants and co-ordinate system; small-difference colorimetry; material standards of color; one-dimensional color scales; and general methods.

**Radiological Laboratory** is a new four-page folder available from New York University College of Engineering. It describes problems of using radioactive isotopes and discusses hazards, special equipment required, how atomic wastes are handled, and gives data on the design and operation of such a laboratory. Copies of the illustrated folder may be obtained for 10 cents from V. W. Phalen, Bureau of Public Information, New York University, New York 53, N. Y.

**Audio Anthology** has been published recently by Radio Magazines, Inc., 342 Madison Ave., New York 17, N.Y., 124 pages. Paper-bound edition is priced at \$2.00 and hard-cover edition at \$3.00.

A compilation of material, designed for the high-fidelity enthusiast, the volume is published to satisfy the demand for articles which appeared in early issues of *Audio Engineering*. Since *Audio Engineering* first made its appearance in 1947, those interested in audio as a hobby, or professionally, have found in it many articles of interest.

## Traveling-Wave Tubes by J. R. Pierce

Published (1950) by D. Van Nostrand Co., Inc. 250 Fourth Ave., New York, N. Y. 223 pages+34-page appendices.+2-page index+ix pages. 137 figures. 6½×9½. \$4.50.

All radio engineers interested in learning of progress towards full utilization of microwaves will welcome the appearance of this book on "Traveling-Wave Tubes" written by John Pierce, so well known for his many important contributions to the art of electron dynamics. This book is the first on traveling-wave tubes and develops the theory of operation of such tubes in considerable detail. It is a "must" for every tube engineer concerned with problems of design and development of traveling-wave and related tubes.

The author points out in the introduction (Chapter I) that in addition to high gain (over 30 db) and the operating frequency range from 200 to 40,000 Mc, the most important characteristic of the traveling-wave tube is its extraordinary bandwidth (3 to 1). This makes the traveling-wave tube (TWT) a most welcome addition to the microwave tube family, since it promises to utilize fully the possibilities of broad-band service where multichannel distortionless transmission is required.

As indicated by the author in the preface to his book, "the emphasis is on general aspects of various wave circuits and their coupling to and interaction with electron flow." After a brief description of the basic form of the TWT (single helix), Chapter II is devoted to the small-signal theory in which the excitation of the circuit by the electron beam and its effect on the circuit are analyzed and the characteristic equation determining the propagation constant is derived. Concepts of gain parameter and beam impedance are introduced and the tube gain is expressed in these terms. Chapter III is devoted to the analysis of wave propagation on a helix and the theory of coupled helices. Detailed mathematical treatment for this and other chapters is contained in the appendices. Chapter IV deals with filter-type circuits where field solutions are derived for simple geometries and more complex circuits are treated as lumped-circuit analogues to which filter network analysis applies.

In Chapter V the effectiveness of circuit interaction with electron beams is analyzed and the desired properties of circuits, such as high impedance, are expressed in terms of stored energy, phase and group velocities, etc. Chapter VI is devoted to a general treatment of waveguiding circuits where their wave propagating properties are described in terms of normal modes. Chapter VII makes use of the generalized circuit representation and derives the characteristic equation for the propagation constant in a more general way, taking into account forces due to space-charge fields. Effects of circuit attenuation and of space-charge forces on tube gain, obtained from solutions of the characteristic equation for special cases, are discussed in Chapter VIII.

In Chapter IX initial boundary conditions and the effect of discontinuities in waveguiding structure are discussed in such a manner that the over-all tube gain can be evaluated. In Chapter X the effect of noise

resulting from random fluctuation in the electron stream of the TWT is analyzed, making use of solutions for the one dimensional case of space-charge flow. Partition noise and noise due to transverse motions of electrons are also considered. In Chapter XI it is shown that backward waves do not produce effective gain, but only wave interference effects. Chapter XII discusses high-level operation and gives estimates of efficiency which are expressed in terms of parameters used in the small signal theory. In Chapter XIII transverse motions of electrons are considered and the gain formula is derived for the transverse type of TWT. Chapter XIV deals with field solutions where the effect of variation of the rf fields over the cross section of the electron beam is considered. The two final Chapters (XV and XVI) are devoted to the introduction to the theory of magnetron and double-stream amplifier tubes.

The reviewer believes that John Pierce's book on "Traveling-Wave Tubes" represents a most competent treatment of the TWT theory, containing the generally accepted basic theory which will serve as a very useful guide to tube designers, and some material of more controversial nature which will inspire many fruitful discussions and will stimulate creative thought.

ANDREW HAEFF  
Naval Research Laboratory,  
Washington, D. C.

## Advances in Electronics, Vol. II, Edited by L. Marton

Published (1950) by Academic Press, Inc., 125 E. 23 St., New York, N. Y. 362 pages+15-page index+ix pages. 160 figures. 9×6. \$7.60.

This second volume, like Volume I published in 1948, consists of survey articles on a variety of topics in the general field of electronics, each article written by a different expert. Volume I contained 10 such articles; the present volume contains 8, ranging between 30 and 60 pages in length. Additional volumes are to be published to provide the electronics specialist with yearbooks resembling, in some respects, those published in another field under the title, "Reports on Progress in Physics" (Physical Society of London).

It is inevitable that the general approach, literary style, scope, and method of treatment in such a collection will have as many variations as the book has articles. There is also a danger that such a book may have a selection of topics so diverse that, in this day of extreme specialization, it will be rare for one individual to be greatly interested in more than a few of the topics. In this respect, Volume II is markedly better than Volume I; the topics are largely in the field of physical electronics and, in the main, are more closely related than were those of Volume I.

The general utility of a collection of partially unrelated survey papers in a bound volume would be subject to question if existing technical journals were already covering the field competently. It may safely be said that, in the electronics and radio engineering sphere, we have not had the careful and periodic selection and preparation of survey papers comparable to those of the *Reviews of Modern Physics*. For this

reason, "Advances in Electronics" fulfills a definite need. On the other hand, in comparing with publication in journal form, the latter assumes adequate attention by the abstracting and bibliographical services, and at the same time, achieves wider initial circulation. It may be that important and carefully worked out papers, when published in book form, will be overlooked by future workers because the content is not clearly established by the title of the book. The remedy is to include articles from these compendia in the standard abstract lists but, to this reviewer's knowledge, it is not being done.

The first four articles of Volume II are related to the cathode-ray art. Dr. Hilary Moss, of Electronic Tubes, Ltd., Reading, England, opens the book with "Cathode Ray Tube Progress in the Past Decade With Special Reference to Manufacture and Design." The title is a reasonably accurate description but, it should be added, the emphasis is on British practice. The treatment appears to be intended for the practical but well-informed engineer and knowledge of the terminology is presupposed. The author is never hesitant to express a definite opinion on controversial points and the paper is largely based on his own experiences. For example, there are 29 references (which is certainly not a large number for a survey) and 10 are to the author's own work. The design information is largely empirical rather than basic, and is a little too limited to a particular type of gun design to survey many years of design changes. The article by Professor Grivet of the University of Paris, "Electron Lenses," is of quite different character. It is basic in nature and intended for the man interested in fundamentals; it is well documented by 100 references, and emphasizes high-performance lenses, as is quite fitting today. This article will be of value to the cathode-ray tube designer, but even more so to the electron microscope worker and the research man using electron optical principles in other devices. Dr. G. Liebmann, of Associated Electrical Industries, Ltd., Berkshire, England, has contributed another well-written basic article on field plotting and electron-ray tracing which, with its 129 bibliographical references, is an excellent complement to the lens papers. These two, together, cover the field of electron optics perhaps as well as could possibly be done in the allotted space. The fourth of the surveys devoted to the cathode-ray art is by Dr. G. F. Garlick of the University of Birmingham, England, and is entitled, "Cathodoluminescence." This review of the physical properties of phosphors which are useful in cathode-ray tubes is one of the shorter items in the book, but is a concise minimal treatment of cathodoluminescent materials, their characteristics and theoretical interpretations, written from the point of view of a physicist interested in the fundamentals.

The remaining four contributions are not so directly related to each other. Professors Frohlich and Simpson of the University of Liverpool have written "Intrinsic Dielectric Breakdown in Solids." This is a good summary of the experimental phenomena and the underlying theoretical hypotheses; one should not expect, and the author does not

include, lists of materials and their breakdown points, since this type of material is better suited to standard handbooks. Professor Gunnar Hok, of the University of Michigan (up to this point the first author with a U. S. affiliation) has written "The Microwave Magnetron," which is the shortest, and the least complete, survey of the present book. It is not documented in detail and suffers by comparison with the classic and much more elegant 1946 paper of Fisk, Hagstrum, and Hartman. On the other hand, it adds a feature to the book under review which, though it will not alone sell the book, may have value to some who are primarily interested in the other electron tube material. In a less well-known field, Dr. George T. Rado, of the U. S. Naval Research Laboratory, discusses "Ferromagnetic Phenomena at Microwave Frequencies"; this is a thorough and up-to-date treatise which will serve many as an excellent introduction, particularly to current research. The final paper is a good review of microwave absorption by gases, under the title, "Microwave Spectroscopy," and was written by Donald K. Coles of the Westinghouse Research Laboratories. The paper includes a long bibliography and a comprehensive presentation of the important topics in the field.

To sum up the entire book, it is a compilation of selected subjects which will be of particular value to the research worker and to the man outside the realm of research who wishes to keep up with fundamental progress in physical electronics.

E. W. HEROLD  
Radio Corporation of America  
Princeton, N. J.

### Radio and Television Mathematics by Bernhard Fischer

Published (1949) by the Macmillan Company, 60 Fifth Ave., New York, N. Y. 440 pages + 12-pages index + 2-page mathematical index + xviii pages + 147 figures. 5½ × 8. \$6.00.

This is the first edition of a handbook on problems and solutions intended "to serve as a guide and reference book for the practical man, as a collection of problems for instructors, and as a review for those who want to acquire a rapid practical skill in solving problems in preparation for radio license examinations given by the United States Federal Communications Commission." In the opinion of the reviewer, this is a book which would be of no value to the practicing engineer or to the engineering student, and that the author does not intend it to be.

The mathematical prerequisite is a working knowledge of algebra and trigonometry. The author states in the preface, "There is no danger of telling the beginner too much. Electronics has become too universal a field, universal in its all-embracing application and universal in the type of persons working in it. No longer can teachers afford to forget that the obvious may be inconceivable to one unfamiliar with the tools of mathematics." In following this idea, the book consists mostly of the statement of the problem and the formula to be used; then the numerical substitution, and the various steps of computation are given. No statement is

made as to whether the formula can be used generally or, if not, what the limitations of it are. The derivations of only a few formulas are given, and some of the problems are solved using network theory.

The text is made up of four sections. The first and most important section covers the following topics: Circuit Components; Direct-Current Circuits; Alternating-Current Circuits; Vacuum-Tube Fundamentals; Amplifiers; Oscillators; Transmitters; Receivers; Power Supplies; Antenna and Transmission Lines; Television; Measurements; and Industrial and Control Circuits. The second section deals with problems for further practice and consists of 312 problems (and answers) of the type worked out in detail in the first section. Some important tools for radio mathematics are the subject of the third section, in which the uses of the powers of 10 in calculating, the *j*-operator, and polar vectors are explained. There is also a brief discussion of the slide rule. The final section is made up of formulas used in the preceding detailed problems, and miscellaneous tables.

From the viewpoint of the practical radioman the information given under each topic is reasonably complete, except in the following cases. Radio-frequency and intermediate-frequency amplifiers, and FM detectors are not dealt with in the topic on Receivers. In the case of Power Supplies, no discussion was given on three-phase supplies, voltage doubling circuits, the various high-voltage supplies used in television, and bridge rectifiers; regulated supplies were not discussed except for single VR tube supplies. Again, in dealing with the subject of Television no mention is made of sweep generators, blocking oscillators, phase discriminators, automatic-gain-control circuits, and multivibrators. In covering video amplifiers, no calculation of the phase response of the amplifier is made. Also, in using the method applied to the amplifiers it would be difficult to calculate an unknown circuit component, as no general formulas are given for the video amplifier.

Practically all of the calculations required for the Study Guide of the Federal Communications Commission are covered in the text. Only two problems, the calculation of the effective height of an antenna and the speed of an induction motor, are not included.

The book, in general, is easily readable, but some errors were noted which might cause confusion. Under the topic of direct-current circuits some alternating-current calculations are made. In the case of conversion calculation for beam-power tubes the answers obtained are correct, but the statement is made that "the current conversion factor can be found by using the fundamental law of vacuum tubes that  $I_p \propto E_p^{3/2}$  . . . ." This statement is not correct and probably causes considerable confusion. In a beam-power tube, the plate current is practically independent of the plate voltage over a considerable range. The reason the conversion factor is correct in the problem is:

$$I_{\text{space current}} = K \left( E_g + \frac{E_{ac}}{\mu_{sc}} \right)^{3/2},$$

and the plate current and the screen current are each a certain fraction of the total space

current; therefore,

$$I_P + K_1 \left( E_g + \frac{E_{ac}}{\mu_{sc}} \right)^{3/2}$$

and

$$I_{sc} = K_2 \left( E_g + \frac{E_{ac}}{\mu_{sc}} \right)^{3/2}.$$

The new screen and grid voltages are  $E'_{sc} = F_e E_{sc}$  and  $E'_g = E_g E_g$ , where  $F_e$  is the conversion factor. Substituting these voltages in the equation for plate current,

$$\begin{aligned} I_P' &= K_1 \left( F_e E_g + \frac{F_e E_{sc}}{\mu_{sc}} \right)^{3/2} \\ &= \left[ K_1 \left( E_g + \frac{E_{sc}}{\mu_{sc}} \right) \right] F_e^{3/2} \\ &= I_P F_e^{3/2}. \end{aligned}$$

The same procedure would apply for the screen current. There are a few other problems which are incorrect.

On the whole the extent of the material covered in this book is quite impressive. It would probably be useful to the beginner and to the practical radioman who does not know how to make quantitative calculations.

WILLIAM L. BEHREND  
RCA Laboratories  
Princeton, N. J.

### Television for Radiomen by Edward M. Noll

Published (1949) by the Macmillan Company, 60 Fifth Avenue, New York, N. Y. 573 pages + 7-page index + xii pages. 456 figures. 9½ × 6. \$7.00.

This book starts with excellent instructions on how to study a book with the objective of a home study course in mind. It is not so much an equipment designers text as it is a review of present-day television system details. It includes very little abstract theory and design engineering, but (assuming an understanding of ordinary radio circuits) explains in practical terms the principles of television transmission and reception equipment. It uses enough of the essential mathematics to enable a technician working with such apparatus to make necessary adjustments.

A modern television receiver is handled part by part, showing the particular applications of the theory to understand, operate, adjust, or repair these items. The information on commercial equipment is up-to-date, and is well arranged for classroom teaching, home study work, as a text in trade and technical schools, and as a reference book on present-day techniques in TV receivers. The book can thus serve as a basis for those who desire to advance into transmitter operation or into the engineering end of the art. It opens the door for the ambitious radioman to become adept in servicing, giving him the detailed knowledge that permits him to diagnose receiver difficulties. All examples are taken from typical commercial receivers, and all circuit arrangements found in practice seems to be covered. Finally, the book provides "easy reading" for those who do not have an instructor at hand to interpret the meanings.

RALPH R. BATCHER  
Douglaston, L. I.,  
N. Y.

# Institute Committees—1950

## EXECUTIVE

R. F. Guy, *Chairman*  
D. B. Sinclair, *Vice-Chairman*  
Haraden Pratt, *Secretary*

S. L. Bailey      A. N. Goldsmith  
W. R. G. Baker      J. W. McRae

## BOARD OF EDITORS

A. N. Goldsmith, *Chairman*

F. W. Albertson	S. S. Mackeown
J. S. Allen	Nathan Marchand
William Bachman	E. D. McArthur
G. M. K. Baker	Knox McIlwain
W. L. Barrow	J. W. McRae
R. R. Batcher	L. A. Meacham
B. B. Bauer	G. F. Metcalf
R. M. Bowie	R. A. Miller
Ralph Bown	E. L. Nelson
R. S. Burnap	D. O. North
O. H. Caldwell	H. F. Olson
C. W. Carnahan	R. M. Page
C. C. Chambers	H. O. Peterson
L. M. Clement	G. W. Pickard
J. D. Cobine	Haraden Pratt
M. G. Crosby	C. A. Priest
R. B. Dome	J. R. Ragazzini
W. G. Dow	Simon Ramo
E. W. Engstrom	H. J. Reich
W. L. Everitt	J. D. Reid
W. G. H. Finch	F. X. Rettenmeyer
D. G. Fink	P. C. Sandretto
H. C. Forbes	H. H. Scott
I. A. Getting	S. W. Seeley
A. W. Graf	V. W. Sherman
F. W. Grover	L. C. Smyby
D. B. Harris	C. E. Smith
L. B. Headrick	J. A. Stratton
E. W. Herold	W. C. Tinus
J. K. Hilliard	K. S. Van Dyke
J. A. Hutcheson	E. K. Van Tassel
J. K. Johnson	H. A. Wheeler
L. F. Jones	J. R. Whinnery
H. S. Knowles	W. C. White
J. D. Kraus	L. E. Whittemore
J. B. H. Kuper	Jerome Wiesner
J. L. Lawson	G. W. Willard
D. G. Little	I. G. Wolff
F. B. Llewellyn	V. K. Zworykin

## AWARDS

M. G. Crosby, *Chairman*

S. L. Bailey	A. V. Loughren
D. E. Chambers	J. W. McRae
Keith Henney	B. E. Shackelford
F. S. Howes	Karl Spangenberg

## EDUCATION

H. J. Reich, *Chairman*

R. G. Anthes	J. M. Pettit
R. E. Beam	W. H. Pickering
L. J. Black	Albert Preisman
C. C. Chambers	L. R. Quarles
R. M. Fano	J. R. Ragazzini
G. H. Fett	J. D. Ryder
R. A. Galbraith	R. P. Siskind
A. W. Graf	W. R. Smith
A. E. Harrison	Karl Spangenberg
R. W. Hickman	F. R. Stansel
G. B. Hoadley	F. E. Terman
G. L. Hollander	L. A. Ware
A. H. Howell	Ernst Weber
A. W. Melloh	L. E. Williams
P. H. Nelson	A. H. Wing, Jr.

Irving Wolff

## CONSTITUTION AND LAWS

I. S. Coggeshall, *Chairman*

R. D. Chipp	F. B. Llewellyn
R. A. Heising	Haraden Pratt
	H. R. Zeamans

## PROFESSIONAL GROUPS

W. R. G. Baker, *Chairman*

S. L. Bailey	R. A. Heising
I. S. Coggeshall	J. V. L. Hogan
W. H. Doherty	J. D. Reid
W. L. Everitt	B. E. Shackelford

(Professional Group Chairmen  
Ex-Officio)

## TELLERS

J. L. Callahan, *Chairman*

E. J. Isbister	A. A. McKenzie
----------------	----------------

## ADMISSIONS

P. S. Christaldi, *Chairman*

H. S. Bennett	C. A. Hachemeister
C. M. Burrill	L. N. Hatfield
A. G. Clavier	R. E. Mathes
H. P. Corwith	H. S. Moncton
F. M. Deelhake	J. H. Moore
T. M. Ferrill, Jr.	G. B. Riley
	H. C. Vance

## POLICY DEVELOPMENT

Haraden Pratt, *Chairman*

K. C. Black	T. A. Hunter
A. V. Eastman	F. B. Llewellyn
W. L. Everitt	J. D. Reid
W. R. Hewlett	B. E. Shackelford
J. V. L. Hogan	J. A. Stratton
	W. N. Tuttle

## SECTIONS

J. F. Jordan, *Chairman*

Robert Broding	George Rappaport
R. A. Heising	J. E. Shepherd
T. A. Hunter	E. T. Sherwood
H. I. Metz	L. C. Sigmon
C. A. Norris	R. N. White

(Section Chairmen Ex-Officio)

## NOMINATIONS

R. A. Heising, *Chairman*

Ralph Bown	W. L. Everitt
Robert Broding	Keith Henney
Melville Eastham	W. R. Hewlett
	F. H. R. Pounsett

## MEMBERSHIP

T. H. Clark, *Chairman*

F. W. Albertson	E. C. Jordan
J. E. Brown	W. A. Knoop
A. B. Chamberlain	O. I. Lewis
J. B. Coleman	F. B. Llewellyn
H. C. Forbes	F. L. Marx
G. W. Fyler	W. P. Short
V. M. Graham	D. B. Smith
R. N. Harmon	F. E. Terman
G. L. Hollander	W. C. White

(Chairmen of Section Membership  
Committees Ex-Officio)

## PUBLIC RELATIONS

R. R. Batcher, *Chairman*

S. L. Bailey	W. L. Everitt
W. R. G. Baker	E. K. Gannett
E. L. Bragdon	R. A. Hackbusch
O. H. Caldwell	Keith Henney
W. C. Copp	T. R. Kennedy
O. E. Dunlap, Jr.	M. B. Sleeper
	Lewis Winner

## FINANCE

W. R. G. Baker, *Chairman*

S. L. Bailey	I. S. Coggeshall
	D. B. Sinclair, ex-officio

## PAPERS REVIEW

G. F. Metcalf, *Chairman*

H. A. Affel	M. T. Lebenbaum
P. H. Betts	C. V. Litton
F. J. Bingley	W. P. Mason
D. S. Bond	R. E. Mathes
Kenneth Bullington	H. F. Mayer
H. A. Chinn	L. L. Merrill
J. K. Clapp	H. R. Mimno
S. B. Cohn	F. L. Moseley
J. M. Constable	G. G. Muller
M. G. Crosby	A. F. Murray
F. W. Cunningham	J. R. Nelson
A. R. D'heedene	K. A. Norton
M. J. DiToro	Ernest Pappenfus
H. D. Doolittle	H. W. Parker
O. S. Duffendack	L. J. Peters
R. D. Duncan, Jr.	A. P. G. Peterson
E. H. Felix	W. H. Pickering
V. H. Fraenckel	A. F. Pomeroy
R. L. Freeman	Albert Preisman
Paul Fritschel	T. H. Rogers
E. G. Fubini	H. E. Roys
Stanford Goldman	J. D. Ryder
W. M. Goodall	M. W. Scheldorf
G. L. Haller	Samuel Seely
O. B. Hanson	Harner Selvidge
A. E. Harrison	R. E. Shelby
T. J. Henry	J. E. Smith
C. N. Hoyler	R. L. Snyder
P. K. Hudson	E. E. Spitzer
D. L. Jaffe	J. R. Steen
Hans Jaffe	G. C. Sziklai
Henry Jasik	H. P. Thomas
D. C. Kalbfell	Bertram Trevor
A. G. Kandoian	W. N. Tuttle
Martin Katzin	Dayton Ulrey
J. G. Kreer, Jr.	S. N. Van Voorhis
Emile Labin	J. R. Weiner
J. J. Lamb	M. S. Wheeler
V. D. Landon	R. M. Wilmotte
	H. R. Zeamans

# Technical Committees, May 1, 1950-May 1, 1951

## ANNUAL REVIEW

R. R. Batcher, *Chairman*  
R. T. Hamlett, *Vice-Chairman*

H. E. Allen	P. C. Sandretto
George M. Brown	P. F. Shea
Donald H. Castle	R. E. Shelby
Trevor Clark	H. E. Singleton
John Crawford	W. N. Tuttle
E. L. Harder	G. L. Van Deusen
J. V. L. Hogan	K. S. Van Dyke
E. C. Jordan	D. E. Watts
R. M. Mitchell	Ernst Weber
W. J. Poch	H. P. Westman
A. F. Pomeroy	I. R. Weir
George Rappaport	H. W. Wells
Nathaniel Rochester	L. E. Whittemore
A. L. Samuel	R. E. Zenner

## ANTENNAS AND WAVE GUIDES

A. G. Fox, *Chairman*  
Sidney Frankel, *Vice-Chairman*

T. M. Bloomer	D. C. Ports
P. S. Carter	H. J. Riblet
L. J. Chu	V. H. Rumsey
W. S. Dutterra	M. W. Scheldorf
J. E. Eaton	S. A. Schelkunoff
Henry Jasik	S. Sensiper
E. C. Jordan	J. P. Shauklin
M. L. Kales	George Sinclair
O. E. Kienow	P. H. Smith
W. E. Kock	R. N. Sorea
	L. C. Van Atta

## AUDIO TECHNIQUES COMMITTEE

R. A. Miller, *Chairman*  
H. H. Scott, *Vice-Chairman*

O. L. Angevine, Jr.	W. W. Dean
H. W. Augustadt	J. A. Green
L. L. Beranek	H. D. Harris
W. L. Black	J. K. Hilliard
H. Burris-Meyer	F. L. Hopper
C. A. Cady	D. E. Maxwell
D. H. Castle	Ralph Schlegel
E. J. Content	W. E. Stewart
A. N. Curtis	R. T. Van Niman

## CIRCUITS

W. N. Tuttle, *Chairman*  
C. H. Page, *Vice-Chairman*

W. R. Bennett	E. A. Guillemin
J. G. Brainerd	W. H. Huggins
Cledo Brunetti	Herbert Krauss
A. R. D'heeden	John Linvill
R. L. Dietzold	P. F. Ordung
W. L. Everitt	E. H. Perkins
R. M. Foster	Wolcott Smith
Stanford Goldman	Ernst Weber
	J. R. Weiner

## ELECTROACOUSTICS

E. S. Seeley, *Chairman*  
B. B. Bauer, *Vice-Chairman*

H. E. Allen	F. V. Hunt
Eginhard Dietze	W. F. Meeker
M. J. Di Toro	H. F. Olson
Martin Greenspan	Vincent Salmon
E. C. Gregg	P. S. Veneklasen
H. F. Hopkins	John Volkman

## ELECTRON TUBES AND SOLID-STATE DEVICES

L. S. Nergaard, *Chairman*  
A. L. Samuel, *Vice-Chairman*

R. S. Burnap	W. G. Dow
J. W. Clark	C. E. Fav

A. M. Glover	J. A. Morton
J. E. Gorham	I. E. Mouromtseff
J. W. Greer	G. D. O'Neill
L. B. Headrick	O. W. Pike
L. A. Hendricks	P. A. Redhead
E. C. Homer	H. J. Reich
S. B. Ingram	A. C. Rockwood
R. B. Janes	R. M. Ryder
S. J. Koch	R. W. Slinkman
R. L. McCreary	H. L. Thorson
	C. M. Wheeler

## ELECTRONIC COMPUTERS

J. W. Forrester, *Chairman*  
Nathaniel Rochester, *Vice-Chairman*

D. R. Brown	G. W. Patterson
R. B. Elbourne	J. A. Rajchman
E. L. Harder	Robert Serrell
John Howard	R. L. Snyder
E. Lakatos	J. R. Weiner
B. R. Lester	C. F. West
G. D. McCann	W. D. Woo

## FACSIMILE

J. V. L. Hogan, *Chairman*  
C. J. Young, *Vice-Chairman*

James Barnes	F. A. Hester
Henry Burkhard	Pierre Mertz
J. J. Callahan	N. A. Nelson
A. G. Cooley	Hugh C. Ressler
	R. J. Wise

## INDUSTRIAL ELECTRONICS

D. E. Watts, *Chairman*  
John Dalke, *Vice-Chairman*

G. P. Bosomworth	H. W. Parker
J. M. Cage	S. I. Rambo
E. W. Chapin	Walther Richter
J. E. Eiselein	W. C. Rudd
C. W. Frick	C. F. Spitzer
G. W. Klingaman	E. H. Schulz
H. R. Meahl	W. R. Thurston
Eugene Mittelmann	R. S. Tucker
P. E. Ohmart	M. P. Vore
	Julius Weinberger

## MEASUREMENTS AND INSTRUMENTATION

Ernst Weber, *Chairman*  
F. J. Gaffney, *Vice-Chairman*

Wilson Aull	W. J. Mayo-Wells
Carl C. Chambers	G. A. Morton
P. S. Christaldi	C. D. Owens
John Dalke	A. P. G. Peterson
G. L. Fredendall	J. G. Reid, Jr.
C. W. Frick	J. R. Steen
W. D. George	N. H. Taylor
E. J. Green	R. S. Tucker

## MOBILE COMMUNICATIONS

F. T. Budelman, *Chairman*  
Alexander Whitney, *Vice-Chairman*

G. M. Brown	D. E. Noble
D. B. Harris	J. C. O'Brien
C. M. Heiden	David Talley
C. N. Kimball, Jr.	George Teomney
	R. W. Tuttle

## MODULATION SYSTEMS

Bertram Trevor, *Chairman*  
W. G. Tuller, *Vice-Chairman*

F. L. Burroughs	J. G. Kreer, Jr.
W. J. Cunningham	E. R. Kretzmer
L. A. W. East	V. D. Landon
A. C. Goodnow	L. A. Meacham
D. D. Grieg	Dale Pollack
D. M. Hill	H. E. Singleton

## NAVIGATION AIDS

P. C. Sandretto, *Chairman*  
H. R. Mimno, *Vice-Chairman*  
C. J. Hirsch, *Vice-Chairman*

W. B. Burgess	Wayne Mason
Henri Busignies	K. A. Norton
G. C. Comstock	W. M. Richardson
Harry Davis	L. M. Sherer
D. G. Fink	Ben Thompson
	R. R. Welsh

## PIEZOELECTRIC CRYSTALS

K. S. Van Dyke, *Chairman*  
R. A. Sykes, *Vice-Chairman*

C. F. Baldwin	Clifford Frondel
W. L. Bond	B. S. George
W. G. Cady	Edward Gerber
J. K. Clapp	Hans Jaffe
W. A. Edson	W. P. Mason
	P. L. Smith

## RADIO TRANSMITTERS

H. Tanck, *Chairman*  
A. E. Kerwien, *Vice-Chairman*

E. L. Adams	J. B. Heffelfinger
T. J. Boerner	J. B. Knox
Paul Breen	L. A. Looney
M. R. Briggs	J. F. McDonald
H. R. Butler	John Ruston
L. T. Findley	Berthold Sheffield
Harold Goldberg	Harry Smith
	I. R. Weir

## RECEIVERS

R. F. Shea, *Chairman*  
J. D. Reid, *Vice-Chairman*

J. Avins	I. E. Lempert
G. L. Beers	C. R. Miner
J. E. Brown	Garrard Mountjoy
W. F. Cotter	J. F. Myers
R. T. Cox	J. M. Pettit
A. R. Hodges	F. H. R. Pounsett
K. W. Jarvis	S. W. Seeley
J. K. Johnson	S. C. Spielman
	W. O. Swinyard

## SOUND RECORDING AND REPRODUCING

H. E. Roys, *Chairman*  
A. W. Friend, *Vice-Chairman*

S. J. Begun	Everett Miller
M. S. Corrington	A. R. Morgan
George Graham	A. P. G. Peterson
G. P. Hixenbaugh	Harry Schecter
F. L. Hopper	R. A. Schlegel
E. W. Kellogg	Ward Shepard, Jr.
R. A. Lynn	Lincoln Thompson
J. Z. Menard	C. F. West
	R. E. Zenner

## STANDARDS

J. G. Brainerd, *Chairman*  
L. G. Cumming, *Vice-Chairman*  
A. G. Jensen, *Vice-Chairman*

M. W. Baldwin, Jr.	A. F. Pomeroy
R. R. Batcher	George Rappaport
H. G. Booker	H. E. Roys
F. T. Budelman	P. C. Sandretto
W. G. Cady	E. S. Seeley
P. S. Carter	R. F. Shea
J. W. Forrester	J. R. Steen
A. G. Fox	H. Tanck
R. A. Hackbusch	Bertram Trevor
J. V. L. Hogan	W. N. Tuttle
J. E. Keister	L. C. Van Atta
E. A. Laport	K. S. Van Dyke
Wayne Mason	D. E. Watts
R. A. Miller	Ernst Weber
L. S. Nergaard	L. E. Whittemore

## SYMBOLS

A. F. Pomeroy, *Chairman*  
 A. G. Clavier, *Vice-Chairman*  
 K. E. Anspach  
 C. R. Burrows  
 H. F. Dart  
 E. T. Dickey  
 W. J. Everts  
 W. A. Ford  
 R. T. Haviland  
 O. T. Laube  
 C. D. Mitchell  
 C. Neitzert  
 M. B. Reed  
 Duane Roller  
 A. L. Samuel  
 E. W. Schafer  
 M. S. Smith  
 W. F. Snyder  
 H. P. Westman

## TELEVISION SYSTEMS

A. G. Jensen, *Chairman*  
 R. E. Shelby, *Vice-Chairman*  
 W. F. Bailey  
 M. W. Baldwin, Jr.  
 R. M. Bowie  
 A. H. Brolly  
 J. E. Brown  
 K. A. Chittick

C. G. Fick  
 D. G. Fink  
 C. J. Franks  
 P. C. Goldmark  
 R. N. Harmon  
 J. L. Hollis  
 I. J. Karr  
 R. D. Kell  
 P. J. Larsen  
 H. T. Lyman

Leonard Mautner  
 J. Minter  
 J. H. Mulligan, Jr.  
 A. F. Murray  
 J. A. Ouimet  
 D. W. Pugsley  
 David Smith  
 M. E. Strieby  
 A. Talamini  
 Norman Young, Jr.

L. W. Morrison, Jr.  
 R. S. O'Brien  
 C. G. Pierce  
 J. F. Wiggin

## WAVE PROPAGATION

H. G. Booker, *Chairman*  
 H. W. Wells, *Vice-Chairman*

E. W. Allen, Jr.  
 S. L. Bailey  
 J. E. Boyd  
 C. R. Burrows  
 T. J. Carroll  
 A. B. Crawford  
 A. E. Cullum, Jr.  
 W. S. Duttera  
 E. G. Fubini  
 I. H. Gerks  
 M. C. Gray  
 D. E. Kerr  
 J. E. Keto  
 Morris Kline  
 M. S. Morgan  
 K. A. Norton  
 H. O. Peterson  
 George Sinclair  
 Newbern Smith  
 A. W. Straiton  
 A. H. Waynick  
 J. W. Wright  
 R. L. Smith-Rose

## VIDEO TECHNIQUES

J. E. Keister, *Chairman*  
 W. J. Poch, *Vice-Chairman*  
 M. W. Baldwin, Jr.  
 A. J. Baracket  
 P. F. Brown  
 R. H. Daugherty, Jr.  
 V. J. Duke  
 G. L. Fredendall  
 R. L. Garman  
 L. L. Lewis

## Special Committees

## PROFESSIONAL RECOGNITION

G. B. Hoadley, *Chairman*  
 C. C. Chambers  
 Harry Dart  
 W. E. Donovan  
 C. M. Edwards

## EDITORIAL ADMINISTRATIVE

A. N. Goldsmith, *Editor, Chairman*  
 G. M. K. Baker  
 H. S. Black  
 R. S. Burnap  
 E. F. Carter  
 R. L. Dietzold  
 (alternate)  
 A. V. Haefl  
 E. W. Herold  
 F. B. Llewellyn  
 Knox McIlwain  
 Donald McNicol  
 K. A. Norton  
 (alternate)  
 Haraden Pratt  
 J. R. Ragazzini  
 F. X. Rettenmeyer  
 (alternate)  
 Newbern Smith  
 G. C. Sziklai  
 (alternate)  
 Ernst Weber  
 H. A. Wheeler  
 L. E. Whittemore  
 Harold Zahl

ARMED FORCES LIAISON  
COMMITTEE

G. W. Bailey, *Chairman*

## INSTITUTE REPRESENTATIVES IN COLLEGES—1950\*

\*Agricultural and Mechanical College of Texas: Tom Prickett, Jr.  
 \*Akron, University of: P. C. Smith  
 \*Alabama Polytechnic Institute: G. H. Saunders  
 \*Alberta, University of: J. W. Porteous  
 \*Arizona, University of: H. E. Stewart  
 \*Arkansas, University of: G. H. Scott  
 British Columbia, University of: H. J. MacLeod  
 \*Brooklyn, Polytechnic Institute of: A. B. Giordano  
 \*Bucknell, University of: R. C. Walker  
 \*California Institute of Technology: W. H. Pickering  
 \*California State Polytechnic College: Clarence Radius  
 \*California, University of: L. J. Black  
 California, University of at Los Angeles: E. F. King  
 Carleton College: G. R. Love

\*Carnegie Institute of Technology: E. M. Williams  
 \*Case Institute of Technology: J. D. Johannesen  
 Cincinnati, University of: A. B. Bereskin  
 \*Clarkson College of Technology: F. A. Record  
 \*Colorado, University of: H. W. Boehmer  
 \*Columbia University: J. R. Ragazzini  
 \*Connecticut, University of: C. W. Schultz  
 \*Cooper Union: J. B. Sherman  
 \*Cornell University: True McLean  
 Dartmouth College: M. G. Morgan  
 \*Dayton, University of: Appointment later  
 \*Delaware, University of: H. S. Bueche  
 Denver University of: Herbert Reno  
 \*Detroit, University of: Appointment later  
 Drexel Institute of Technology: R. T. Zern  
 Duke University: W. J. Seeley  
 Evansville College: J. F. Sears  
 \*Fenn College: K. S. Sherman  
 \*Florida, University of: H. A. Owen  
 \*George Washington University: W. S. Carley

\*Georgia Institute of Technology: M. A. Honnell  
 Harvard University: E. L. Chaffee  
 Idaho, University of: appointment later  
 \*Illinois Institute of Technology: G. F. Levy  
 \*Illinois, University of: E. C. Jordan  
 \*Iowa, State University of: L. A. Ware  
 \*Iowa State College: G. A. Richardson  
 \*John Carroll University: J. L. Hunter  
 Johns Hopkins University: Ferdinand Hamburger, Jr.  
 \*Kansas State College: J. E. Wolfe  
 Kansas, University of: D. G. Wilson  
 \*Kentucky, University of: H. W. Farris  
 \*Lafayette College: F. W. Smith  
 Lawrence Institute of Technology: H. L. Byerlay  
 \*Lehigh University: D. E. Mode  
 Louisiana State University: Appointment later  
 \*Louisville, University of: Appointment later  
 \*Maine, University of: W. J. Creamer, Jr.  
 \*Manhattan College: T. P. Canavan

\* Colleges with approved Student Branches

Manitoba, University of: R. G. Anthes  
 \*Marquette University: G. A. Frater  
 \*Maryland, University of: George Corcoran  
 \*Massachusetts Institute of Technology: E. A. Guillemin, W. H. Radford  
 McGill University: F. S. Howes  
 \*Miami, University of: Appointment later  
 \*Michigan College of Mining and Technology: R. J. Jones  
 \*Michigan State College: M. D. Rogers  
 \*Michigan, University of: L. N. Holland  
 \*Minnesota, University of: O. A. Becklund  
 \*Mississippi State College: Appointment later  
 \*Missouri School of Mines and Metallurgy: G. G. Skitek  
 \*Missouri, University of: G. V. Lago  
 \*Nebraska, University of: Charles Rook  
 Nevada, University of: I. J. Sandorf  
 \*Newark College of Engineering: M. E. Zaret  
 New Hampshire, University of: A. L. Winn  
 \*New Mexico College of Agriculture and Mechanic Arts: H. A. Brown  
 \*New Mexico, University of: Tom Martin  
 \*New York, College of the City of: Harold Wolf  
 \*New York University: Philip Greenstein  
 \*North Carolina State College: G. B. Hoadley  
 \*North Dakota, University of: Clifford Thomforde  
 \*Northeastern University: G. E. Pihl  
 Northwestern University: A. H. Wing, Jr.  
 Notre Dame, University of: H. E. Ellithorn

\*Ohio State University: E. M. Boone  
 \*Oklahoma Agricultural and Mechanical College: H. T. Fristoe  
 Oklahoma, University of: C. L. Farrar  
 \*Oregon State College: A. L. Albert  
 \*Pennsylvania State College: C. R. Ammerman  
 \*Pennsylvania, University of: C. C. Chambers  
 \*Pittsburgh, University of: J. F. Pierce  
 \*Pratt Institute: David Vitrogo  
 \*Princeton University: N. W. Mather  
 \*Purdue University: R. P. Siskind  
 Queens University: H. H. Stewart  
 \*Rensselaer Polytechnic Institute: H. D. Harris  
 \*Rhode Island State College: J. L. Hummer  
 Rice Institute: C. R. Wischmeyer  
 Rose Polytechnic Institute: H. A. Moench  
 \*Rutgers University: J. L. Potter  
 \*San Diego State College: D. C. Kalbfell  
 Santa Clara, University of: W. J. Warren  
 \*San Jose State College: Harry Engwicht  
 \*Seattle University: Appointment later  
 \*South Carolina University of: L. R. Wever  
 \*South Dakota School of Mines and Technology: Appointment later  
 \*Southern California, University of: G. W. Reynolds  
 \*Southern Methodist University: E. J. O'Brien  
 \*Stanford University: J. M. Pettit  
 \*St. Louis University: G. E. Dreifke  
 \*Stevens Institute of Technology: A. C. Gilmore, Jr.

\*Syracuse University: R. P. Lett  
 \*Tennessee, University of: E. D. Shipley  
 Texas Technological College: Appointment later  
 \*Texas, University of: C. M. Crain  
 \*Toledo, University of: R. E. Weeber  
 \*Toronto, University of: George Sinclair  
 \*Tufts College: A. H. Howell  
 \*Tulane University: M. E. Forsman  
 Union College (N.Y.): R. B. Russ  
 Union College (Nebr.): M. D. Hare  
 United States Military Academy: F. K. Nichols  
 United States Naval Post Graduate School: G. R. Giet  
 \*Utah State Agricultural College: Clayton Clark  
 \*Utah, University of: O. C. Haycock  
 \*Virginia Polytechnic Institute: R. R. Wright  
 \*Virginia, University of: J. C. Mace  
 Washington, State College of: appointment later  
 \*Washington, University of: V. L. Palmer  
 Washington University: S. H. Van Wambeek  
 \*Wayne University: M. B. Scherba  
 Western Ontario, University of: E. H. Tull  
 West Virginia University: R. C. Colwell  
 \*Wisconsin, University of: Glenn Koehler  
 Witwatersrand, University of: G. R. Bozzoli  
 \*Worcester Polytechnic Institute: H. H. Newell  
 \*Wyoming, University of: W. M. Mallory  
 \*Yale University: H. J. Reich

## INSTITUTE REPRESENTATIVES ON OTHER BODIES—1950

American Association for the Advancement of Science: J. C. Jensen  
 American Documentation Institute: J. H. Dellinger  
 ASA Standards Council: J. G. Brainerd, L. G. Cumming (alternate)  
 ASA Conference of Staff Executives: G. W. Bailey, L. G. Cumming (alternate)  
 ASA Electrical Standards Committee: L. G. Cumming, F. B. Llewellyn, E. A. LaPort  
 ASA Sectional Committee (C16) on Radio: V. M. Graham (Chairman), J. G. Brainerd, L. G. Cumming, J. J. Farrell  
 ASA Sectional Committee (C18) on Specifications for Dry Cells and Batteries: H. M. Turner  
 ASA Sectional Committee (C39) on Electrical Measuring Instruments: Wilson Aull, Jr.  
 ASA Sectional Committee (C42) on Definitions of Electrical Terms: J. G. Brainerd, A. G. Jensen, Haraden Pratt, H. A. Wheeler  
 ASA Subcommittee (C42.1) on General Terms: J. C. Jensen  
 ASA Subcommittee (C42.6) on Electrical Instruments: J. H. Miller  
 ASA Subcommittee (C42.13) on Communications: J. C. Schelleng  
 ASA Subcommittee (C42.14) on Electronics: R. S. Burnap  
 ASA Sectional Committee (C60) on Standardization of Electron Tubes: L. S. Nergaard, C. E. Fay  
 ASA Sectional Committee (C61) on Electric

and Magnetic Magnitudes and Units: J. H. Dellinger  
 ASA Sectional Committee (C63) on Radio-Electrical Co-ordination: C. C. Chambers  
 ASA Sectional Committee (C67) on Standardization of Voltages—Preferred Voltages—100 Volts and Under: A. F. Van Dyck  
 ASA Sectional Committee (Z10) on Letter Symbols and Abbreviations for Science and Engineering: H. M. Turner, A. F. Pomeroy (alternate)  
 ASA Sectional Committee (Z14) on Standards for Drawings and Drafting Room Practices: Austin Bailey  
 ASA Sectional Committee (Z17) on Preferred Numbers: A. F. Van Dyck  
 ASA Sectional Committee (Z24) on Acoustical Measurements and Terminology: Eginhard Dietze, H. F. Olson  
 ASA Sectional Committee (Z32) on Graphical Symbols and Abbreviations for Use on Drawings: Austin Bailey, A. F. Pomeroy (alternate)  
 ASA Subcommittee (Z32.9) on Communication Symbols: H. M. Turner  
 ASA Sectional Committee (Z57) on Standards for Sound Recording: S. J. Begun  
 ASA Subcommittee (Z57.1) on Magnetic Recording: W. J. Morlock  
 ASA Sectional Committee (Z58) on Standardization of Optics: E. D. Goodale, L. G. Cumming (alternate)

ASME Glossary Review Board: W. R. G. Baker  
 IRE, AIEE Committee on Noise Definitions: Stanford Goldman, Jerry Minter, Claude Shannon  
 IRE, RMA Problem No. 2: S. L. Bailey  
 Joint EEI, NEMA, RMA Co-ordination Committee on Radio Reception: C. E. Brigham  
 Joint IRE, ASA C42 Definitions Co-ordinating Committee: C. L. Dawes, A. G. Jensen, J. C. Schelleng  
 Joint IRE, AIEE, NEMA Co-ordination Committee on Commercial Induction and Dielectric Heating Apparatus: D. E. Watts  
 Joint RMA-IRE Co-ordination Committee: R. F. Guy, Keith Henny  
 Joint Technical Advisory Committee: R. F. Guy  
 National Electronics Conference Board of Directors: J. D. Reid  
 National Research Council, Division of Engineering and Research: F. B. Llewellyn  
 URSI (International Scientific Radio Union) Executive Committee: C. M. Jansky, Jr.  
 U. S. National Committee, Advisers on Electrical Measuring Instruments: Melville Eastham, H. L. Olesen  
 U. S. National Committee, Advisers on Symbols: L. E. Whitmore, J. W. Horton  
 U. S. National Committee of the International Electrotechnical Commission: L. G. Cumming, E. A. LaPort, F. B. Llewellyn

# Abstracts and References

Prepared by the National Physical Laboratory, Teddington, England, Published by Arrangement  
with the Department of Scientific and Industrial Research, England,  
and *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, and not to the IRE.

Acoustics and Audio Frequencies.....	1235
Antennas and Transmission Lines.....	1236
Circuits and Circuit Elements.....	1236
General Physics.....	1238
Geophysical and Extraterrestrial Phenomena.....	1239
Location and Aids to Navigation.....	1240
Materials and Subsidiary Techniques..	1240
Measurement and Test Gear.....	1243
Other Applications of Radio and Electronics.....	1244
Propagation of Waves.....	1245
Reception.....	1245
Stations and Communication Systems..	1245
Subsidiary Apparatus.....	1246
Television and Phototelegraphy.....	1246
Transmission.....	1247
Tubes and Thermionics.....	1247
Miscellaneous.....	1248

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

## ACOUSTICS AND AUDIO FREQUENCIES

- 534.232** **2112**  
The Production of Edge Tones by Circular Gas Jets and Plane Laminae—H. von Gierke. (*Z. Angew. Phys.*, vol. 2, pp. 97-106; March, 1950.) An experimental investigation and discussion of the theory of the production of edge tones at a hole or slit in a lamina directly above a jet of gas. Stroboscopic observations of the jets under such conditions indicate that the tones are due to the periodic whirls of the issuing gas. Frequency measurements were made for different lamina spacings and different velocities and diameters of jet. The properties of edge-tone devices as radiators of sound energy are considered.

- 534.26+535.421** **2113**  
The Diffraction of a Plane Wave through a Grating—Miles. (See 2179.)

- 534.26** **2114**  
On Two Complementary Diffraction Problems—A. Storruste and H. Wergeland. (*Phys. Rev.*, vol. 73, pp. 1397-1398; June 1, 1948.) Spheroidal coordinates are introduced for obtaining exact solutions for the diffraction of sound waves by a circular disk and by a hole in an infinite plane screen. The electromagnetic case is to be dealt with later.

- 534.321.9:534.133-14** **2115**  
Theoretical and Experimental Investigations of the Oscillations and the Radiation Resistance of an Ultrasonic Quartz Crystal—G. Bolz. (*Z. Angew. Phys.*, vol. 2, pp. 119-127; March, 1950.) The differential equations are derived for the 3-dimensional motion of a quartz crystal oscillating in a liquid. An approximate solution is given, assuming plane oscillations, and expressions are derived for radiation re-

sistance, amplitude of oscillations, decrement and equivalent circuit values near resonance. The mean value of radiation resistance found by measuring the damping effect on an oscillatory circuit agrees with the theoretical expression. Deviation (generally <10 per cent) of the measured value from the mean is attributed to an unknown form factor, the assumption of plane oscillations being only an approximation to the actual conditions. Values given for equivalent  $L$  and  $C$  are valid for a crystal oscillating in air or in vacuo.

- 534.321.9:534.373+534.22** **2116**  
A New Frequency-Modulation Method for Measuring Ultrasonic Absorption in Liquids—E. Ribchester. (*Nature* (London), vol. 165, p. 970; June 17, 1950.) The output from a rf oscillator, modulated by a sinusoidal af voltage, is applied to a quartz transducer. The af voltage is also passed through a variable-phase-shift network and modulates a second oscillator on a different radio frequency. The output from this oscillator is injected into a frequency changer together with the signal from the receiving transducer in the liquid cell, and the phase-shift network is adjusted for maximum output from the frequency-changer, corresponding to absence of frequency modulation. Results of measurements of absorption in water, using a frequency of 10 Mc, are in close agreement with those obtained by pulse techniques.

- 534.6:621.395.632.11** **2117**  
Acoustical Study of Telephone Bells of the French P.T.T. Administration—In 1325 of July please insert as authors P. Chavasse and R. Lehmann.

- 534.86** **2118**  
Listeners' Sound-Level Preferences: Part 2—T. Somerville and S. F. Brownless. (*BBC Quart.*, vol. 5, pp. 57-64; Spring, 1950.) A description of tests on listeners' preferences for changes in sound level between the end of one program and the beginning of the next. The results are summarized for various program changes; a reduction of 4-5 db is preferred for speech following music, and an increase of about 2 db for music following speech. Part 1: 1266 of 1949.

- 621.395.61+621.395.623.7:597.08** **2119**  
Underwater Sound Technique in the Service of Modern Research—H. Gemperle. (*Radio Tech.* (Vienna), vol. 26, pp. 235-239; May, 1950.) Description and theory of pressurized transmitter and microphone units for depths up to 30 m. The frequency range 30 cps-20 kc is covered with good efficiency.

- 621.395.623.7** **2120**  
Study of Different Acoustic Baffles for Loudspeakers—T. S. Korn. (*Toute la Radio*, vol. 17, pp. 183-186; June, 1950.) Analysis of general theory of sound radiation from loud speakers and discussion of different basic types of baffle. The simplest effective construction for even response is the base reflex baffle consisting of a closed cabinet lined with absorbent material, with an orifice beneath the loudspeaker. The labyrinth type baffle gives a similar response curve.

- 621.395.625.3:621.3.018.78†** **2121**  
Over-all Frequency Characteristic in Magnetic Recording—P. E. Axon. (*BBC Quart.*, vol. 5, pp. 46-53; Spring, 1950.) Frequency distortion in magnetic recording is discussed and illustrated by calculated and measured characteristic curves. The frequency characteristic of the record on the tape can be determined independently of the reproducing-head characteristic by using a reproducing head with a wide gap. The magnitude of self-demagnetization in commercial tapes can be evaluated by this method, and the results obtained can be used in conjunction with the calculated characteristic of a normal reproducing head to predict the response of magnetic recording equipment.

- 621.395.665.1** **2122**  
The Electrodynamical Volume Expander—(*Radio Tech. Dig.* (Franc), vol. 4, pp. 116-124 and 177-181; April and June, 1950.) Based on an article in Italian by Zanarini. The principles of a simple and economical expansion circuit are described in which the moving-coil loudspeaker serves as the expander, the current through the field coil being controlled in accordance with the amplitude of the af signal applied to the moving coil. No distortion is introduced by this method and the amplifier chain needs no modification. The basic circuit consists of a transformer with primary across the moving coil; the output from the secondary is rectified and applied across a parallel RC circuit to the grid and cathode of a pentode which has the field coil in its anode circuit. Developments of the circuit are discussed and practical considerations are summarized.

- 621.396.645.029.3** **2123**  
A.F. Amplifiers—R. Besson. (*Toute la Radio*, vol. 17, pp. 167-171 and 187-191, 199; May and June, 1950.) Three classes of amplifier are considered: (a) high-quality; (b) a compromise of quality and cost; and (c) the cheapest with satisfactory performance. Design considerations and details of ancillary circuits are discussed. A 4.5-w and 10-w amplifier of class (b) are treated as examples and simple calculations made to determine their design specifications.

Complete circuit diagrams and notes on construction and testing are included.

## ANTENNAS AND TRANSMISSION LINES

621.315.212:621.317.336 2124  
Study of the Impedance Irregularities of Coaxial Cables by Oscillographic Observation of Pulse Echoes—P. Herreng and J. Ville. (*Ann. Télécommun.*, vol. 3, pp. 317–331; October, 1948.) See 2162 of 1948 and 142 of 1949 (Counault and Herreng).

621.392 2125  
Phase Distortion in Feeders: Effect of Mismatching on Long Lines—L. Lewin, J. J. Muller, and R. Basard. (*Wireless Eng.*, vol. 27, pp. 143–145; May, 1950.) A fresh approach to the problem of transmitting FM signals along feeder lines is made by considering the system as supporting a series of progressive waves of various instantaneous frequencies and reflected within the confines of the feeder. The production of phase distortion and harmonics is analyzed and an example of transmission along a waveguide is worked out.

621.392 2126  
Study of a System of Two Parallel Wires by the Method of Hallén—V. Belevitch. (*HF* (Brussels), no. 6, pp. 163–168; 1950.) An integral equation, analogous to that of Hallén for an antenna, is established for a bifilar line. The first-approximation solution of this equation corresponds with the results given by the classic theory of transmission lines. The second approximation takes account of end effects and of radiation. The expression obtained for the radiated power coincides with that derived by C. Manneback by another method.

621.392.26† 2127  
The Field Generated by an Arbitrary Current Distribution within a Waveguide—J. J. Freeman. (*Bur. Stand. Jour. Res.*, vol. 44, pp. 193–198; February, 1950.) Formulas are derived for waveguides of rectangular, circular, and coaxial cross section by extending expressions previously obtained for cavities (2762 of 1948) to the case of infinite length. As a check on the formulas it is shown that the value given for the field from an axially directed dipole within a circular guide reduces to the free-space value when the radius becomes infinite.

621.392.26† 2128  
Reflection Cancellation in Waveguides—L. Lewin. (*Elec. Commun.*, vol. 27, pp. 48–55; March, 1950.) Reprint. See 3037 of 1949.

621.392.26†:621.396.67 2129  
Slotted Waveguides and Their Applications as Aerials: Part 1—J. Ortusi and G. Boissinot. (*Ann. Radioélec.*, vol. 5, pp. 94–108; April, 1950.) Theory is given based on the elementary concept of the reflection and transmission coefficients of a slot. From consideration of a single slot in (a) an infinite conducting plane, and (b) a waveguide wall, general formulas are derived for the radiation from a waveguide having  $n$  equal and equidistant slots. The relations are applied to waveguides with  $\lambda/2$  and with  $\lambda/4$  spacing of slots, which radiate respectively with constant amplitude and with exponentially decreasing amplitude. Formulas which determine the conditions of use of slotted waveguides and their operating bandwidth are deduced.

621.392.5 2130  
Note on Artificial Delay Lines—G. Martin. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 1645–1647; May 8, 1950.) A method of calculation more exact than hitherto used is developed for a line consisting of a helically wound wire on an insulated cylindrical metal core, by considering the planar transformation in which the system is cut lengthwise and opened out.

The method applies particularly to the case when the dielectric layer is thin and the pitch of the winding and the core curvature are small, and permits evaluation and correction of non-uniformity effects at the end of the line. Measurements by various methods agree with the calculated results to within a few per cent.

621.396.67 2131  
The Principle of Conservation of Energy and Kottler's Formulae—J. Ortusi and J. C. Simon. (*Ann. Radioélec.*, vol. 5, pp. 67–73; April, 1950.) Application of Kottler's formulas to the radiation from a waveguide aperture shows that, considering a waveguide of infinite length, the law of energy conservation and Maxwell's equations are satisfied simultaneously only by an existing field. For an arbitrary field to satisfy the energy condition, the reflected wave must be introduced. In a waveguide of rectangular cross section the reflected wave tends to a zero value as the frequency is increased without limit. A rigorous proof of this is given.

621.396.67 2132  
An All-Band Mobile Antenna System—S. S. Perry. (*QST*, vol. 34, pp. 16–18; June, 1950.) Details of a combination covering the wavelength range 2–160 m. For wavelengths of 2, 6, and 10 m a  $\lambda/4$  whip antenna mounted high on the vehicle is used. For longer waves an antenna center-loaded with a multitapped coil is substituted. This serves as a  $\lambda/4$  antenna for 10.7-m wavelength when all the coil is short-circuited, and resonates at 3.5 Mc (85.7 m) when about 160 turns of the coil are in use. For wavelengths up to 160 m a second loading coil is added in series.

621.396.671 2133  
Indoor Measurement of Microwave Antenna Radiation Patterns by Means of a Metal Lens—G. A. Wootton, R. B. Borts, and J. A. Carruthers. (*Jour. Appl. Phys.*, vol. 21, pp. 428–430; May, 1950.) Laboratory measurements were made using a metal lens to render the spherical wave-front plane, the method being similar to that used in optical diffraction measurements. Comparison of results with those obtained by standard field methods showed little loss of accuracy for angles between  $-15^\circ$  and  $+15^\circ$  for horns of aperture width 32 cm, using a lens of aperture 110 cm and a wavelength of 3.2 cm. A stepped lens was found unsatisfactory, probably because of diffraction caused by the steps.

621.396.671 2134  
Input Impedance of Horizontal Dipole Aerials at Low Heights above the Ground—R. F. Proctor. (*Proc. IEE* (London), vol. 97, pp. 188–190; May, 1950.) "The large increase in the radiation resistance of a horizontal dipole at low heights above the ground, predicted by the mathematical analysis of Sommerfeld and Renner, is confirmed by actual measurement for the case of an essentially pure dielectric ground. Measurements carried out above a series of conducting mats of increasing mesh size corroborate the supposition that the large increase in the radiation resistance at low heights is caused by a corresponding increase in the energy radiated downwards into the dielectric ground."

621.396.671 2135  
Gain of Aerial Systems—P. M. Woodward and J. D. Lawson. (*Wireless Eng.*, vol. 27, p. 166; May, 1950.) Comment on 3057 of 1949 and 820 of May (Bell).

621.396.671.011.2 2136  
Mutual Impedance and Self-Impedance of Coupled Parallel Aerials—B. J. Starkey and E. Fitch. (*Proc. IEE* (London), vol. 97, pp. 129–137; May, 1950.) Existing formulas for the mutual impedance of parallel antennas are ap-

proximate or based on simplifying assumptions. To check these formulas, the authors measured the front/back ratios of the radiation patterns of 2-element arrays, confirming and extending the observations of McPetrie & Saxton (1450 of 1946). The use of "intrinsic impedance," based on a simple transmission-line calculation, for the self-impedance of each antenna resulted in poor agreement between theory and observation. By introducing resistive losses and end-capacitance in the lines, quite good agreement was obtained.

## CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.2 [negative] 2137  
Negative-Resistance Characteristics—A. W. Keen. (*Wireless Eng.*, vol. 27, pp. 175–179; June, 1950.) Negative resistance characteristics as encountered in practice are qualitatively surveyed, and the distinction between the characteristics resulting respectively from voltage and current control of feedback is emphasized. A method of building up a particular characteristic by synthesis of positive and negative elements is described, the basic positive element being represented very closely by the characteristic of a hard diode. The concept of a "negative diode" is introduced, and an equivalent circuit composed of positive and negative diodes is shown for a triode and discussed also for a tetrode.

621.314.2:621.396.645.2 2138  
Bifilar I.F. Coils—S. R. Scheiner. (*Electronics*, vol. 23, pp. 104–107; June, 1950.) The use of transformers with the two coils wound together for coupling the various stages of stagger-tuned if amplifiers renders coupling capacitors unnecessary. The low time constant of the grid circuit gives improved signal-to-noise ratio and the rf choke normally used to feed the anode of the final if tube can be omitted. Detailed analysis of circuits including bifilar transformers is presented and their advantages and limitations are discussed.

621.314.3† 2139  
Magnetic Amplifiers of the Balance-Detector Type—W. A. Geyger. (*Elec. Eng.*, vol. 69, p. 459; May, 1950.) Summary of 1950 AIEE Winter General Meeting paper.

621.316.726.078.3:538.569.4.029.64 2140  
Servo Theory Applied to Frequency Stabilization with Spectral Lines—W. D. Hershberger and L. E. Norton. (*Jour. Frank. Inst.*, vol. 249, pp. 359–366; May, 1950.) Comparison between the frequency of a gas-absorption line and the frequency to be stabilized, or a frequency simply related to it, is made by a sweep-oscillator technique. A pulse is generated each time the resonance frequency is passed through. The resulting voltages are applied to a suitable discriminator, the output from which is fed to a servo system controlling the frequency. The limitations and performance of the circuits involved are discussed. A long-term frequency stability within 1 part in  $10^7$  has been realized.

621.316.8:621.396.822 2141  
Noise in Current-Carrying Ohmic Conductors—B. Meltzer. (*Phil. Mag.*, vol. 41, pp. 393–398; April, 1950.) The mean-square noise current in a conductor which is in temperature equilibrium and carries a current  $I$ , is calculated on the assumption that the current is due to elementary carriers of charge  $e$ . The value obtained for the spectral density is  $\{(2kT/R) + [(2kT/R)^2 + (2eI)^2]^{1/2}\}$  where  $R$ ,  $T$ ,  $k$  are respectively resistance, absolute temperature, and Boltzmann's constant. The interpretation of this result is that, over and above the ordinary thermal Nyquist noise, there is present an additional noise, of which the spectral density varies in square-law fashion according to  $(e^2R/kT)I^2$  for small values of  $I$ , and in linear

fashion according to  $\{(-2kT/R) + 2eI\}$  for large values of  $I$ . These predictions are compared with Bernamont's experimental results for thin metallic conductors, and the discrepancies are discussed.

#### 621.316.86 2142

**Thermistors: Part 1. Static Characteristics**—O. J. M. Smith. (*Rev. Sci. Instr.*, vol. 21, pp. 344–350; April, 1950.) "The fundamental Thermistor equations are presented as functions only of electrical quantities and constants, and independent of temperature measurements. The universal curves allow one to predict the static characteristics of thermistors as circuit elements. The two most significant characteristics are the power coefficient of resistance, and the negative incremental resistance. These can be computed, or read from the universal curves, if the resistance and four parametric constants are known. These five values can be determined from six convenient measurements: the resistance and change of resistance with current at very low current, at the maximum voltage point, and at maximum current." Part 2: 2143 below.

#### 621.316.86 2143

**Thermistors: Part 2. Dynamic Characteristics**—O. J. M. Smith. (*Rev. Sci. Instr.*, vol. 21, pp. 351–356; April, 1950.) The thermistor sinusoidal and transient response characteristics are presented as functions of the dissipation constant and incremental resistance. Equations are given for time constant, critical frequency, equivalent inductance, and  $Q$ . The construction of a low-frequency oscillator is discussed, as well as possible nonlinear-filter applications. Part 2: 2142 above.

#### 621.318.572 2144

**Decade Pulse Counter for Geiger-Müller Tubes**—B. Åström. (*Rev. Sci. Instr.*, vol. 21, pp. 323–326; April, 1950.) Full circuit details of a counter with the following features:—direct indication; easily reset to zero; all resistors and capacitors with 10-per cent tolerances; very reliable and insensitive to power-line fluctuations; highly sensitive; components easily accessible; high-voltage supply for G-M tube capable of supplying enough current for a quenching circuit of the Neher-Pickering type.

#### 621.318.572 2145

**Pentode Counting or Control Ring**—B. L. Moore. (*Rev. Sci. Instr.*, vol. 21, pp. 337–338; April, 1950.) Ten pentodes, Type 6AS6, are used in a very stable counting or control ring whose operation is essentially independent of the incoming pulses, and which does not require careful selection of tubes.

#### 621.392 2146

**A Useful Network Procedure**—S. C. Dunn. (*Wireless Eng.*, vol. 27, pp. 171–175; June, 1950.) The method establishes a relation between the parallel and the star connection of quadripoles. Any number of networks in parallel, with common generator and load, can be replaced by a set in star connecting, having a common central load, but each having an independent generator. The method is applied to study the steady-state and transient responses of a parallel-T RC network. The use of the method for obtaining the input impedance of multichannel networks is demonstrated, and an example is given of a prism-like system of resistor wires which may be reduced to a network of the separate-generator type.

#### 621.392.4 2147

**Transformation of a Lumped-Constant Two-Pole Network to a Dissipative Transmission Line**—A. Millán. (*Rev. Telecommunicación* (Madrid), vol. 4, pp. 2–14; March, 1949.) The short-circuited dissipative transmission line whose admittance simulates, over a particular frequency band, the input admittance of a

given lumped-parameter circuit, is determined by developing the appropriate irrational function in terms of Mittag-Leffler partial fractions. The graph of the resultant approximately rational function is compared with that of the rational function for the given network, and the transmission-line parameters deduced. Alternatively the lumped-parameter circuit is simulated by a lossless line terminated by a dissipative rod. An illustrative example is calculated.

#### 621.392.43 2148

**Matching of Amplifiers and Centimetre-Wave Autooscillators over a Wide Band of Frequencies**—M. Denis. (*Ann. Radioélec.*, vol. 5, pp. 74–88; April, 1950. *Onde Élec.*, vol. 30, pp. 271–284; June, 1950.) Typical examples are given of maximum amplifier efficiency and minimum distortion occurring at different frequencies. The advantages of ideal matching in this sense are discussed with particular reference to the klystron amplifier. Matching circuits are described which are derived from band-pass filters or from natural or artificial delay lines. The phase-changing element necessary for an asymmetrical quadripole matching device is analogous to the circuits used in the microwave field. The extension of these principles to self-oscillators and amplifier chains is discussed, the resulting improvement in efficiency and reduction of the number of stages making the sacrifice of simplicity worth while.

#### 621.392.43 2149

**Method of Matching a Feeder Line for a Given Frequency**—J. Paris. (*Radio Franç.*, no. 6, pp. 4–8; June, 1950.) Formulas are derived for finding the positions of the Sterba clip and its short-circuiting link for matching a 2-wire feeder of negligible loss to an antenna or other terminating load. The use of the Sterba diagram and of a cartesian diagram is illustrated by numerical examples.

#### 621.392.43 2150

**Graphical Methods for Impedance Matching**—A. F. Huerta. (*Rev. Telecommunicación* (Madrid), vol. 4, pp. 15–22; March, 1949.) With special reference to stub matching.

#### 621.392.5 2151

**A Graphical Approach to the Synthesis of General Insertion Attenuation Functions**—J. M. Linke. (*Proc. IEE* (London), vol. 97, pp. 179–187; May, 1950.) An investigation of methods of constructing the insertion transfer function of a network as a rational function meeting prescribed requirements. The insertion modulus function has been determined approximately by earlier methods of limited usefulness. A new semigraphical method is described, using template curves on log-log paper. These are selected to fit the required zeros and poles of the numerator and denominator polynomials and are added graphically to give the approximate attenuation curve shape. The method is unsuitable for networks having sharp cutoff regions.

#### 621.392.5:621.315.212:621.397.743 2152

**Equalization of Coaxial Lines**—K. E. Gould. (*Elec. Eng.*, vol. 69, p. 390; May, 1950.) Summary of 1949 AIEE Fall General Meeting paper. Arrangements used for television service are discussed; the measured gain and delay-distortion characteristics for the Washington/Chicago circuit are shown graphically.

#### 621.392.52 2153

**Design of Dissipative Band-Pass Filters Producing Desired Exact Amplitude-Frequency Characteristics**—M. Dishal. (*Elec. Commun.*, vol. 27, pp. 56–81; March, 1950.) Reprint. See 3369 of 1949.

#### 621.392.52 2154

**Calculation of Transients in Electrical**

**Filters**—T. Laurent. (*Rev. Telecommunicación* (Madrid), vol. 4, pp. 17–28; December, 1948.) Application of the frequency-transformation method noted in 1339 of 1938 to obtain expressions for transients in low-pass and band-pass filters. Translated from *Tekniska Meddelander från Kungl. Telegrafstyrelsen*, no. 2, August, 1947.

#### 621.392.52:539.2 2155

**Physical Basis of the Wave Filter**—Salt-peter. (See 2197.)

#### 621.392.52:621.397.82 2156

**The Latest Techniques for the Elimination of Ham TVI**—P. S. Rand. (*CQ*, vol. 6, pp. 9–12, 61; June, 1950.) Description of the construction and correct method of fitting low-pass filters in antenna leads for harmonic suppression, together with methods of interference suppression for ignition systems.

#### 621.395.665.1 2157

**The Electrodynamical [volume] Expander**—(See 2122.)

#### 621.395+621.396].813 2158

**Distortion**—Please note that, in this and subsequent issues, the UDC number used will be 621.3.018.78† instead of 621.395.813 and 621.396.813 used hitherto.

#### 621.396.6:061.4 2159

**Trends in Components**—(*Wireless World*, vol. 56, pp. 229–235; June, 1950.) Short account of selected exhibits at the private exhibition organized by the Radio and Electronic Component Manufacturers' Federation and held in London, April 17–19, 1950. A list of the names and addresses of exhibitors is appended.

#### 621.396.611.1:517.942.932 2160

**Relaxation Oscillations**—LaSalle. (See 2261.)

#### 621.396.611.21 2161

**Loading Quartz Crystals to Improve Performance**—(*Tech. Bull. Nat. Bur. Stand.*, vol. 34, pp. 59–62; May, 1950.) Experimental investigation by L. T. Sogn. Examination by means of a probe electrode of the frequency and activity distribution at the surface of a crystal plate led to the detection of differences in thickness of  $<20 \mu$ . Variations in surface activity and frequency for various plates, loaded and unloaded, and during temperature runs from  $-60^\circ\text{C}$  to  $+90^\circ\text{C}$  are described. In many cases loading has proved very beneficial.

#### 621.396.611.21.012.8 2162

**X-Cut Quartz Crystal**—F. M. Leslie. (*Wireless Eng.*, vol. 27, pp. 180–181; June, 1950.) "A new equivalent circuit for a crystal driving a mechanical load is devised in terms of an electrical transmission line instead of the usual lumped components of inductance and capacitance."

#### 621.396.615.11 2163

**The Calculation of RC Oscillators**—G. Isay. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 40, pp. 509–517; August 6, 1949. In German.) The basic design principles of RC oscillators are outlined and illustrated by an example in which the calculation for a negative-feedback amplifier is based on that for two simple amplifier stages with negative current feedback, the first stage having a relatively low value of cathode resistance so that feedback to the second stage is reduced. The sharp resonance of the amplifier using a Wien-bridge type of feedback circuit, and two methods of amplitude limiting by means of thermistors, are discussed. Phase-shift oscillators are briefly considered, with particular application to the design of a 3-phase generator.

621.396.615.142.2.018.424 2164

**A Wide Range 600-7000 Mc/s Local Oscillator**—P. Janis. (*Tele-Tech*, vol. 9, pp. 22-24; April, 1950.) The frequency range 600-7000 Mc is covered on cw by two reflex klystrons, Types 6BL6 and 6BM6, used in conjunction with an adjustable cavity resonator consisting of a short-circuited coaxial line. Design and construction details are given. For pulsed outputs, Types SD-1103 and SD-1104 reflex klystrons have been developed; these have similar operating characteristics to the cw tubes, but have a low-voltage control electrode for taking the input pulses.

621.396.615.17/18 2165

**A Tuned Plate Multivibrator**—A. E. Johanson. (*Bell Lab. Rec.*, vol. 28, pp. 208-212; May, 1950.) The difficulty of obtaining short-duration pulses with the conventional multivibrator circuit is overcome by replacing the anode resistors by parallel combinations of varistor rectifiers and tuned circuits, whose resonance frequency then controls the timing. Pulses of duration estimated at 0.02  $\mu$ s have been produced. A further modification is described for frequency division at high frequencies, e.g., from 18.432 Mc to 8 kc in four steps.

621.396.615.18 2166

**Frequency Division with Phase-Shift Oscillators**—C. R. Schmidt. (*Electronics*, vol. 23, pp. 111-113; June, 1950.) Divisions by factors as high as 7 are easily obtainable by means of single-tube RC oscillators using standard components requiring only initial adjustment. Dividers operating from crystal- or tuning-fork-controlled oscillators have been constructed to give accurate 60-cps outputs. Divisions by 5, 6, and 7 were used in these designs, the locking range for these factors being adequate to ensure reliable operation. Initial adjustment of single-stage scale-of-ten circuits requires more care because of the restricted locking range.

621.396.619.23 2167

**Non-Linear Distortion in a Cowan Modulator**—V. Belevitch. (*Wireless Eng.*, vol. 27, pp. 164-165; May, 1950.) Author's reply to 1552 of July (Tucker and Jeynes).

621.396.645 2168

**A New Amplifier, the "Cathode Repeater"**—V. J. Cooper. (*Marconi Rev.*, vol. 13, pp. 72-80; 1950.) The use of a triode with the cathode as the signal input terminal has certain advantages for wide-band amplifiers. Under different conditions its input impedance may be purely resistive or may have a positive or negative susceptance. The circuit described can amplify from a low-impedance input into a low-impedance output, with considerable control of the values of input and output impedance.

621.396.645 2169

**Differential Amplifiers**—D. L. Johnston. (*Wireless Eng.*, vol. 27, p. 194; June, 1950.) Comment on 1636 of August (Parnum).

621.396.645 2170

**Amplifier with Negative-Resistance Load**—D. M. Tombs and M. F. McKenna. (*Wireless Eng.*, vol. 27, pp. 189-193; June, 1950.) Stage gains up to 2.5  $\mu$  (where  $\mu$  is the tube amplification factor) have been obtained with a triode amplifier loaded with a negative-resistance element. The restrictions imposed on the operation of this type of amplifier by considerations of stability, and the limitation of the gain due to the capacitance of the load are discussed.

621.396.645+621.396.621:621.385 2171

**Electronic Valves: Books II-IV. [Book Review]**—(See 2388.)

621.396.645:621.3.015.3].012 2172

**Transient Response Calculation**—J. E. Flood. (*Wireless Eng.*, vol. 27, pp. 182-188; June, 1950.) The response of various types of

$n$ -stage amplifier circuits may be expressed in terms of the Poisson exponential probability summation. Values of this function are already available in tables and charts. Examples are given of its use in analysis of single-stage amplifiers with resistance or LC coupling, and of multi-stage amplifiers with RC or filter coupling or with critically damped inductance compensation.

621.396.645.018.424 2173

**Stagger Gain Calculator**—E. R. Jenkins. (*Tele-Tech*, vol. 9, p. 29; April, 1950.) Graphs are presented for determining the optimum number of stages required in a wide-band amplifier of specified characteristics.

621.396.645.35 2174

**A D. C. Amplifier for Biological Application**—P. O. Bishop and E. J. Harris. (*Rev. Sci. Instr.*, vol. 21, pp. 366-377; April, 1950.) A four-channel high-gain direct-coupled amplifier is described, with an input impedance  $>100$  M $\Omega$  at 10 kc. The frequency response is flat from zero to over 10 kc. Alternative capacitor coupling is provided so that large-amplitude very-low-frequency noise may be eliminated. When direct-coupled, the zero shift is of the order of 100  $\mu$ v peak to peak over a period of 30 minutes.

621.396.645.37:621.526 2175

**Feedback Amplifiers and Servo Systems**—E. E. Ward. (*Wireless Eng.*, vol. 27, pp. 146-153; May, 1950.) The pattern of poles and zeros of the response function is considered as a basis for classifying and studying these devices; this avoids discrimination between the frequency and transient responses, and the Laplace transform is used to develop a broad theory which includes both as special cases. The analysis is restricted to linear systems having lumped constants and one closed loop. A tentative classification of response functions on this basis is tabulated.

621.396.645.371 2176

**Some Considerations in the Design of Negative-Feedback Amplifiers**—W. T. Duerdoth. (*Proc. IEE* (London), vol. 97, pp. 138-155; May, 1950. Discussion, pp. 155-158.) "A brief survey is made of the application of a phase-amplitude theorem, due to Bode, to the design of feedback amplifiers with constant-phase margins, and a graphical method of estimating the phase change at any frequency from a knowledge of the amplitude characteristic is described. The use of constant-phase margin as a design criterion is criticized, and justification is given for a new criterion termed 'stability margin.' Using this criterion, a technique is developed for employing several feedback paths, thus enabling an increased amount of feedback to be connected to circuits of increased complexity. A brief account of a miniature audio amplifier with two feedback paths is included, and a 100-watt amplifier covering 300 cps-108 kc with some 35 db of feedback is described which, although it is 'conditionally' stable, is suitable for use in a multichannel telephony system."

621.392.52 2177

**An Introduction to the Theory and Design of Electric Wave Filters [Book Review]**—F. Scowen. Publisher: Chapman and Hall, London, 2nd edn., 1950, 188 pp. 18s. (*Jour. Brit. I.R.E.*, vol. 10, p. vi; May, 1950.) First published in 1945, this book "has already established itself as a work of major importance" in its field. A new section now added describes Darlington's insertion-loss method of filter design.

## GENERAL PHYSICS

53 2178

**Heaviside's Unpublished Notes**—(*Electrician*, vol. 144, p. 1749; May 26, 1950.) Re-

port of a paper "Some Unpublished Notes of Oliver Heaviside," by H. J. Josephs, given at one of the I.E.E. Heaviside Centenary Commemorative Meetings, May, 1950. The documents in the possession of the IEE show that the fourth volume of "Electromagnetic Theory" was to contain an account of his ideas concerning the transformation of matter and energy on the atomic scale, together with his integration of the electrodynamics of Clerk Maxwell with the thermodynamics of Willard Gibbs. He extended his conception of "energy-tubes" to include their condensation into electronic particles. Other notes deal with the particle-like structure of electromagnetic pulses and the correlation of electromagnetism with gravitation.

534.26+535.421 2179

**The Diffraction of a Plane Wave through a Grating**—J. W. Miles. (*Quart. Appl. Math.*, vol. 7, pp. 45-64; April, 1949.) The problem of diffraction and scattering of a normally incident plane wave of sound by an infinite plane grating consisting of infinitely thin, coplanar, equally spaced strips with parallel edges is solved. The solution is extended to deal with em waves incident on a grating the elements of which are perfectly conducting at rf or perfectly reflecting at optical frequencies. The case when the screen elements are of finite thickness  $t$  is also considered and an approximate solution is obtained including only first-order terms in  $t/\lambda$ .

534.26+535.42]:517.564.4 2180

**The Oblate Spheroidal Wave Functions**—Leitner and Spence. (See 2259.)

535.37:548.0 2181

**On the Mechanism of Phosphorescence in Crystals**—D. Curie. (*Jour. Phys. Radium*, vol. 11, pp. 179-185; April, 1950.) Study of the decay of phosphorescence of long persistence. If the return of electrons to the emission centers is bimolecular their recapture is favored and the decay of phosphorescence is retarded. In certain cases however the general law of decay still holds.

535.42 2182

**Diffraction at a Circular Disk**—W. Braunkbek. (*Z. Phys.* vol. 127, pp. 405-415; March 17, 1950.) The method previously described (2183 below) is applied to the case of a circular disk and to related problems. For the near field at points on the axis a very simple expression is obtained which agrees with the rigorous solution better than the Kirchhoff approximation. Simple formulas are also derived for the distant field. For the transmission coefficient on account of mathematical difficulties no improvement on the Kirchhoff approximation is obtained.

535.42 2183

**New Approximation Method for Diffraction at a Plane Screen**—W. Braunkbek. (*Z. Phys.*, vol. 127, pp. 381-390; March 17, 1950.) Various approximation methods for diffraction of a scalar wave have been proposed, but none give accurate results for small wavelengths (large  $ka$ ), since their series converge very slowly when  $ka$  is large,  $a$  being a linear dimension of the disk or aperture and  $k$  the coefficient in the scalar wave equation  $\Delta u + k^2 u = 0$ . A new approximation method has been developed specially for the case where  $ka \gg 1$ . Numerical comparison with the rigorous solution for a particular case shows that the new method gives considerably better results than the Kirchhoff approximation down to  $ka \approx 1$ .

535.42:538.56 2184

**A More Exact Fresnel-Field Diffraction Relation**—G. A. Woonton. (*Canad. Jour. Res.*, vol. 28, pp. 120-126; March, 1950.) The field close to a radiating aperture is often of interest in radio-optical measurements, but the usual

approximations in the Fresnel-field evaluation are generally invalid in this region. The distant field is given with good accuracy by the Fraunhofer relation, and from it the near Fresnel field is obtained by the Fourier-transform method. The analysis applies to rectangular apertures for which the field distribution can be expressed as the product of separate distributions in the two directions. The result is an integral which has not been evaluated in closed form, but which reduces to Fresnel's integral when the appropriate approximations are made.

535.42:538.56 2185

On the Diffraction of an Electromagnetic Wave through a Plane Screen—J. W. Miles. (*Jour. Appl. Phys.*, vol. 21, p. 468; May, 1950.) Corrections to paper abstracted in 202 of February.

537.217 2186

Vibration Pressure Hypothesis for Electric Repulsion—H. B. Dwight. (*Elec. Eng.*, vol. 69, pp. 397-398; May, 1950.) Similarity between the equation for the net repulsion between moving charges and the equation for the pressure exerted on one moving body by sound emitted from another is adduced in support of Mackaye's theory that electric repulsion is a manifestation of radiation pressure.

537.224 2187

Electrets—W. F. G. Swann. (*Phys. Rev.*, vol. 78, p. 811; June 15, 1950.) An expression is derived for the potential of an electret assuming that it consists of (a) a distribution of semi-permanent polarization which decays with time, and (b) a distribution of surface and volume charge which leaks away and has no relation to the decay of the polarization. Complete mathematical details will be published later.

537.226 2188

The Lorentz Correction in Barium Titanate—J. C. Slater. (*Phys. Rev.*, vol. 78, pp. 748-761; June 15, 1950.) It is assumed that the ferroelectric behavior of  $\text{BaTiO}_3$  arises because the Lorentz correction leads to a vanishing term in the denominator of the expression for the dielectric constant. The Lorentz correction is computed for the actual crystal structure and it is found that the local field near the Ti ion is much greater than previously believed, so that even a small ionic polarization of the Ti ions will lead to ferroelectric effects.

537.311.3 2189

The Electrical Conductivity of Thin Wires—R. B. Dingle. (*Proc. Roy. Soc. A*, vol. 201, pp. 545-560; May 23, 1950.) Simple approximate methods are first given for evaluating the conductivity of films and wires of a size comparable with the mean free path of the conduction electrons. Rigorous theory for a thin wire is then developed to cover both inelastic and elastic collisions of the electrons at the metal surface. Finally Andrew's recent experimental results for a thin mercury wire are fitted to the theoretical curves obtained, and the mean free path evaluated.

537.311.5:621.396.671 2190

Currents on the Surface of an Infinite Cylinder Excited by an Axial Slot—C. H. Papas and R. King. (*Quart. Appl. Math.*, vol. 7, pp. 175-182; July, 1949.) "The distribution of surface-current density is determined on an axially infinite, perfectly conducting cylinder. A constant distribution of electric field across a narrow axial slot is assumed given."

537.525.029.6 2191

Electron Diffusion in a Spherical Cavity—A. D. MacDonald and S. C. Brown. (*Canad. Jour. Res.*, vol. 28, pp. 168-174; March, 1950.) The diffusion equation for electrons in a non-uniform field is solved in terms of the confluent hypergeometric function, and the condition for

breakdown is derived. Measurements of breakdown fields for uniform electric fields enable the effective characteristic diffusion length to be found. Hence theoretical breakdown curves for nonuniform fields are predicted; these are in agreement with experimental results.

537.562:538.561 2192

Radio Emission of Purely Thermal Origin in Ionized Media—J. F. Denisse. (*Jour. Phys. Radium*, vol. 11, pp. 164-171; April, 1950.) Radiation due to "hyperbolic" transitions similar to those occurring in the X-ray and  $\gamma$ -ray bands is studied theoretically by an analysis simplified from that of Kramers. By combining classical and quantum theory, a general expression is obtained for the radiation coefficient which is valid in all cases. The corresponding absorption calculated by application of Kirchhoff's theorem agrees with the value obtained from Lorentz's formula, and hence the influence of the refractive index of the medium on the radiation can be studied. Gyromagnetic radiation is briefly discussed.

537.71 2193

The M.K.S. or Giorgi System of Units—L. H. A. Carr. (*Beama Jour.*, vol. 57, pp. 101-106 and 145-147; April and May, 1950.) A paper given at the IEE, in March, 1950, putting the case for the adoption of this system. Rationalization is discussed as a separate subject in an appendix.

538.3 2194

Electromagnetic Theory—(*Electrician*, vol. 144, p. 1750; May 26, 1950. Report of a paper "An Appreciation of Heaviside's Contribution to Electromagnetic Theory," by W. Jackson, given at one of the IEE Heaviside Centenary Commemorative Meetings, May, 1950. Heaviside's development of a comprehensive theory of line transmission is not easy to discern from his published work, partly because of repetitions and partly on account of the style, though many of his writings are very clear and the chapter on em waves in vol. 1 of "Electromagnetic Theory" is probably the best physical discussion of em wave propagation along telephone lines yet written. Recognition is given to his work in interpreting and extending Maxwell's theory. A partial treatment (published in 1888) of  $E_{01}$ -mode propagation within a cylindrical conducting tube is also noted.

539.172.4:537.311.33 2195

Transmutation-Produced Germanium Semiconductors—J. W. Cleland, K. Lark-Horovitz, and J. C. Pigg. (*Phys. Rev.*, vol. 78, pp. 814-815; June 15, 1950.) Ge exposed to neutron bombardment becomes a  $p$ -type semiconductor due to lattice displacements and to transmutations. The former can be removed by heat treatment and good agreement is then observed for the number of carriers per  $\text{cm}^3$  as calculated from cross sections and as measured by the Hall effect.

539.172.4:537.311.33 2196

Fast Neutron Bombardment Effects in Germanium—J. H. Crawford, Jr. and K. Lark-Horovitz. (*Phys. Rev.*, vol. 78, pp. 815-816; June 15, 1950.)

539.2:621.392.52 2197

Physical Basis of the Wave Filter—J. L. Salpeter. (*Philips Tech. Commun.* (Australia), no. 1, pp. 3-10; 1950.) A simple explanation is given of the action of periodic lumped-reactance filters, and the analogy with the behavior of a row of particles such as atoms in a crystal lattice when irradiated by X rays is demonstrated, the existence of cut-off frequency and stop- or pass-bands being due to the structural discontinuity in both cases. A brief discussion of the 'sone' theory of solids is included. The exponential electrical line is an instance where

the existence of a cutoff frequency is not an indication of discontinuous structure.

501:512.9 2198

Calcolo Tensoriale e Applicazioni [Book Review]—Finzi and Pastori. (See 2268.)

537.226 2199

Time Dependence of Electronic Processes in Dielectrics. Technical Report Reference L/T219 [Book Review]—H. Fröhlich and J. O'Dwyer. Publisher: British Electrical and Allied Industries Research Association, London, 1949, 8 pp., 4s. 6d. (*Beama Jour.*, vol. 57, p. 151; May, 1950.)

537.71 2200

The M.K.S. System of Electrical and Magnetic Units. British Standard No. 1637 [Book Review]—Publisher: British Standards Institution, London, 1950, 8 pp., 2s. (*Wireless Eng.*, vol. 27, pp. 167-168; May, 1950.) A summarized record of decisions taken by the International Electrotechnical Commission, expressing no opinions either for or against the system. The question of rationalization is not discussed.

538.1 2201

Die Maxwell'sche Theorie in Veränderter Formulierung [Book Review]—L. Kneissler. Publisher: Springer Verlag, Vienna, 51 pp. (*Wireless Eng.*, vol. 27, p. 167; May, 1950.) A suggested modification of Maxwell's theory. "By eliminating magnetism and replacing it by the currents that produce it within the material... Maxwell's theory is brought into agreement with the electron theory in its macroscopic form." The book is a development of articles published in *Archiv für Elektrotechnik* in 1940-2.

538.1 2202

Magnetism [Book Review]—D. Shoenberg. Publisher: Sigma Books, London, 1949, 216 pp., 10s. 6d. (*Nature* (London), vol. 166, p. 5; July 1, 1950.) "None of the essential theory of elementary magnetism is omitted, but it appears in a guise less formidable and more convincing than usual." "The book is of a more advanced character than most of the Sigma series and it should be useful to the serious student as well as to the general reader."

#### GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

521.15:538.12 2203

A 4-Dimensional Generalization of Wilson's Hypothesis—A. Papapetrou. (*Phil. Mag.*, vol. 41, pp. 399-404; April, 1950.) "Wilson's hypothesis of electric currents accompanying the rotation of electrically neutral matter (in order to explain the origin of the geomagnetic field) is generalized in a 4-dimensional language. The main consequences of the generalized formulation are discussed."

521.15:538.12 2204

Magnetism and the Rotation of Celestial Bodies—A. E. Benfield. (*Nature* (London), vol. 166, p. 31; July 1, 1950.) The difference between (a) the predictions based on Fermi-Dirac statistics and (b) the requirements of the charge-separation theory of Berlage (1139 of June) is so great that it is hard to see how terrestrial magnetism can be due to this effect to any appreciable extent.

521.15:538.12 2205

The External Gravitational and Electromagnetic Fields of Rotating Bodies—G. L. Clark. (*Proc. Roy. Soc. A*, vol. 201, pp. 488-509; May 23, 1950.) An electromagnetic field of the form proposed by Blackett is shown to be possible, but the analysis does not rule out the possibility that alternative fields may also satisfy the conditions.

523.74 2206  
A Particularly Intense Solar Flare—R. Müller. (*Naturwissenschaften*, vol. 37, p. 137; March, 1950.) Details of observations at Wendelstein Observatory on November 19, 1949.

523.746 2207  
Origin of Sunspots—D. H. Menzel. (*Nature* (London), vol. 166, pp. 31–32; July 1, 1950.) A new theory is suggested, based on the assumption of a solar general magnetic field of polar intensity several gauss, i.e., 10–20 times less than usual estimates. Ionized gas ejected from the poles returns to the surface near the equator, producing intensification of the magnetic field. Collapse occurs when the gravitational acceleration exceeds the buoyancy effect of the magnetic field; sunspots may then result. Details of the theory are to be given later.

523.746 2208  
On the Interpretation of the Sunspot Cycle—W. Grotrian. (*Naturwissenschaften*, vol. 37, p. 163; April, 1950.) Extension of Schwarzschild's calculation of the period of a star to the case of the sun gives values of 9 and 18 years for the solar cycle, when the general field strength is taken as 3,000 gauss. A combination of the Alfvén-Walén (3469 and 3470 of 1945) and Schwarzschild theories offers a conclusive explanation of the sunspot cycle as the period of electromagnetic-hydrodynamic or plasma oscillation of the sun's disk.

550.381 2209  
The Effect of the Compressibility of the Earth on Its Magnetic Field—C. Truesdell. (*Phys. Rev.*, vol. 78, p. 823; June 15, 1950.)

551.51 2210  
The Stratification of the Atmosphere—H. Flohn and R. Penndorf. (*Bull. Amer. Met. Soc.*, vol. 31, pp. 71–77 and 126–130; March and April, 1950.) Suggestions are made for a new classification of atmospheric layers, based on thermal data.

551.510.535 2211  
Theory of the Production of an Ionized Layer in a Non-Isothermal Atmosphere, Neglecting the Earth's Curvature, and Its Application to Experimental Results—J. A. Gledhill and M. E. Szendrei. (*Proc. Phys. Soc.*, vol. 63, pp. 427–445; June 1, 1950.) "A new theory of ionospheric layer formation is developed, in which the temperature is assumed to vary linearly with height. The equations are compared at each step with those obtained by Chapman in his theory of layer formation in an isothermal atmosphere. The equations for the maximum of electron density and its height are also given. The effect of the parameters on the shape of the layer is shown in graphical form. The equations are somewhat complex in form, but an ingenious graphical method has been devised suitable for the application of the theory to results given in the form of a Booker and Seaton parabolic distribution of electronic density with height. By application of the theory to mean hourly values for four summer months in South Africa, values are obtained for the temperature gradient, the temperature at 200 km, and its variation over the middle part of the day. The results obtained are in accordance with previous estimates and offer numerical confirmation of the theory that the atmosphere expands bodily upwards during the middle part of a summer day."

551.510.535:621.396.11 2212  
The Air Force Interest in Sporadic-E Ionization—N. C. Gerson. (*CQ*, vol. 6, pp. 17–19) June, 1950.) An account of cooperation between a band of 425 amateurs in 15 countries and the U. S. Air Force in the study of sporadic-E ionization, particularly regarding the movement of sporadic-E reflection points. The data

obtained may be used to advance theories on the mechanism of sporadic-E effects and to form some conception of diurnal, seasonal, and annual variations.

551.515.4:621.396.9 2213  
Radar Measurements of the Initial Growth of Thunderstorm Precipitation Cells—Hilst and MacDowell. (See 2200.)

551.524.7 2214  
The Temperature Distribution of the Upper Atmosphere over New Mexico—A. Nazarek. (*Bull. Amer. Met. Soc.*, vol. 31, pp. 44–50; February, 1950.) The vertical distribution of temperature was derived, by use of the hydrostatic equation, from pressure data obtained in rocket experiments. The results indicate a steady increase from 210°K at 15 km to 308°K at 55 km, followed by a rapid decrease to 180°K at 85 km and a subsequent increase to 266°K at 120 km. An irregularity occurs at about 100 km, where the temperature decreases slightly, probably owing to dissociation of molecular into atomic oxygen. Seasonal and latitudinal effects are discussed.

551.578.11 2215  
The Size Distribution of Raindrops—A. C. Best. (*Quart. Jour. R. Met. Soc.*, vol. 76, pp. 16–36; January, 1950.) Existing experimental data are examined, together with recent measurements in England and Wales, and the results are presented in a common form. Drop-size distribution is regarded as a function of rate of rainfall only, for simplification, and in many cases the distribution is in agreement with simple formulas. Tables based on these formulas are included showing values of the amount of liquid water per unit volume and of the number and total volume of drops, between certain diameter limits, for various rates of rainfall.

551.594.6+621.315.59 2216  
University Research in Physics—J. A. Teegan. (*Beama Jour.*, vol. 57, pp. 133–138; May, 1950.) A brief survey of work done and in progress at London University on the nature of atmospherics and on the electrical properties of semiconductors.

551.51 2217  
Oscillations of the Earth's Atmosphere [Book Review]—M. V. Wilkes. Publisher: Cambridge University Press, 1949, 74 pp., 12s 6d. (*Proc. Phys. Soc.*, vol. 63, pp. 459–460; June 1, 1950.) "This excellent little volume . . . deals with the oscillations produced daily in the atmosphere by the sun and the moon, as revealed mainly by the . . . daily variations of barometric pressure."

551.510.535:621.396.11 2218  
Short-Wave Radio and the Ionosphere [Book Review]—T. W. Bennington. Publisher: Iliffe and Sons, London, 2nd edn., 138 pp., 10s. 6d. (*Wireless Eng.*, vol. 27, p. 168; May, 1950.) "This book presents the available information on the ionosphere in simple form so that it is usable by nonspecialists in the subject. The approach is nonmathematical."

#### LOCATION AND AIDS TO NAVIGATION

621.396.9 2219  
'La Pipologie': the Art of Interpreting the 'Pips' on a Radar Screen—Vienne and Scotto. (*Radio Tech. Dig.* (Franc), vol. 4, pp. 67–81; April, 1950.) An article on the technique of radar operating, reproduced from a French Service document.

621.396.9:551.515.4 2220  
Radar Measurements of the Initial Growth of Thunderstorm Precipitation Cells—G. R. Hilst and G. P. MacDowell. (*Bull. Amer. Met. Soc.*, vol. 31, pp. 95–99; March, 1950.) Sequence photographs of PPI and RHI (range

height indicator) radar presentations are used for measurement of the vertical and horizontal growth of the storm. Vertical growth is rapid and horizontal growth is uniform.

621.396.9:551.578.1/.4 2221  
The Bright Band—A Phenomenon Associated with Radar Echoes from Falling Rain—J. E. N. Hooper and A. A. Kippax. (*Quart. Jour. R. Met. Soc.*, vol. 76, pp. 125–132; April, 1950.) The echoes obtained with vertical transmission are characterized by an intense echo from a layer at a height slightly below the freezing level. This is more pronounced in frontal rain than in instability showers, and is about 5 to 9 times more intense than the echo from the rain below, allowing for difference in range. Experiments are described which establish the height and thickness of the "bright band" layer. Measurements of the terminal velocity of snowflakes support Ryde's suggestion that the reduction in echo intensity below the bright band is due to the acceleration of water droplets from their velocity as snowflakes to the terminal velocity of raindrops.

621.396.9:551.578.1/.4 2222  
Radar Echoes from Meteorological Precipitation—J. E. N. Hooper and A. A. Kippax. (*Proc. IEE* (London), vol. 97, pp. 89–95; May, 1950. Discussion, pp. 95–97.) A more detailed account of the work referred to in 2221 above.

621.396.9:621.385.832:535.371.07 2223  
Visibility on Cathode-Ray Tube Screens: Positive vs Negative Signals on an Intensity Modulated Scope—Harriman and Williams. (See 2386.)

621.396.933 2224  
How the C.A.A. Flight Tests V.O.R. Ranges—A. E. Jenks. (*Tele-Tech*, vol. 9, pp. 22–24, 56; May, 1950.) The horizon profile as viewed from the antenna is plotted and course distortion in the shadow area forecast. The station error is obtained by the theodolite check, in which a circular course, with the station as center, is navigated and plotted. The line-of-sight cutoff point can be deduced by analysis of course-line records obtained by flying aircraft over known test points and courses. Errors in course indication due to instability of the supply voltage frequency, to the presence of vertical polarization in the transmitted signal and to heading effect are measured.

621.396.933 2225  
Notes on Radiocommunication Systems for Civil Aviation—(Rev. *Telecomunicación* (Madrid), vol. 4, pp. 47–64; March, 1949.) An account of navigational aids considered at the third [British] Commonwealth and Empire Conference on Radio for Civil Aviation.

#### MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 2226  
On Certain Phenomena of Gaseous Emission Observed during the Evacuation of Low-Voltage Incandescent Lamps—L. Dunoyer. (*Le Vide*, vol. 5, pp. 793–806; May, 1950.) Ignition of the filament of 2-v lamps while still connected to the vacuum plant causes an emission of gas which is more rapid and of greater amount, within limits, the longer the time interval between the heating/degassing process and this ignition. This emission was studied with a special thermal micromanometer. It is attributed to adsorbed layers on the walls of the bulb. Immediate sealing after evacuation limits the possibility of subsequent gassing.

535.37 2227  
X-Rays in the Development of Inorganic Phosphors—H. P. Rooksby and A. H. McKeag. (*GEC Jour.*, vol. 17, pp. 89–95; April, 1950.) Greater knowledge of the crystal chemistry of

luminescent materials is required both to explain the phenomena of luminescence and to effect improvements in the materials. Identification, crystalline perfection, the effect of changed matrix composition on fluorescence color, and the role of impurity activators are among the aspects considered in this survey of the ways in which X-ray analysis has been used in the investigation of phosphor characteristics.

535.37 2228

**Trivalent Cations in Fluorescent Zinc Sulfide**—F. A. Kröger and J. Dickhoff. (*Physica's Grav.*, vol. 16, pp. 297–316; March, 1950. In English.) Incorporation of monovalent cations in a lattice consisting of divalent ions is only possible to an appreciable extent when the lack of positive charge resulting from the substitution of a monovalent cation for a divalent one is compensated. This compensation can be effected by a simultaneous incorporation of monovalent anions or of cations of a valency higher than two. An explanation can thus be given of the activation of ZnS by monovalent Ag, Cu, Au and Zn when halogens or trivalent cations are present. Some of the trivalent ions incorporated in this way are found to cause effects of their own (electron traps fluorescence and killing of fluorescence due to the other centers). An atomic model of the centers of fluorescence is described.

535.37 2229

**The Physical Chemistry of the Formation of Fluorescence Centres in ZnS-Cu**—F. A. Kröger and N. W. Smit. (*Physica's Grav.* vol. 16, pp. 317–328; March, 1950. In English.)

538.114 2230

**Magnetic After-Effect Revealed by Means of a Blow**—L. Lliboutry. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 1586–1587; May 3, 1950.) A light blow may be used instead of a small variation of magnetic field for demonstrating this effect in a steel. See also 1660 of August.

538.22 2231

**The a/H Law of Approach [to saturation] and a New Theory of Magnetic Hardness**—L. Néel. (*Jour. Phys. Radium*, vol. 9, pp. 184–192; May, 1948.)

538.221 2232

**Relation between the Anisotropic Constant and the Law of Approach to Saturation of Ferromagnetic Materials**—L. Néel. (*Jour. Phys. Radium*, vol. 9, pp. 193–199; June, 1948.)

538.221 2233

**The Effect of Annealing on the Residual Magnetization of Ferromagnetics**—J. C. Barbier. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 1643–1645; May 8, 1950.)

538.221 2234

**Factors Affecting Magnetic Quality**—R. M. Bozorth. (*Bell Sys. Tech. Jour.*, vol. 29, pp. 251–286; April, 1950.) The substance of Chapter 2 of a forthcoming book entitled "Ferromagnetism." The effects of chemical composition and methods of production and treatment on the final properties of magnetic materials are discussed. Details of relevant manufacturing processes are included, and lists are given of useful materials and their magnetic properties.

538.221 2235

**Thermal Variation of the Coercive Field of Powdered Cobalt**—L. Weil, S. Marfoué, and F. Bertaut. (*Jour. Phys. Radium*, vol. 9, pp. 203–207; June, 1948.)

538.221 2236

**Antiferromagnetism and Intermittent Activation in Chromium—Iron and Vanadium—Iron Alloys**—R. Forrer. (*Compt. Rend. Acad.*

*Sci.* (Paris), vol. 230, pp. 1584–1585; May 3, 1950.)

538.221 2237

**On the Conditions for the Occurrence of Ferromagnetism in Metal Compounds**—J. H. Gislöf. (*Physica's Grav.*, vol. 15, pp. 677–678; September, 1949. In English.) Discussion leading to the conclusion that in ferrites (which possess both octahedron holes and tetrahedron holes) only the metal ions in tetrahedron holes can contribute to the ferromagnetism.

538.221 2238

**On the Ferromagnetism of Ferrites, or Ferromagnetism**—L. Néel. (*Physica's Grav.*, vol. 16, pp. 350–352; March, 1950. In French.) Various experimental results are quoted which indicate that Gislöf's interpretation of the magnetic properties of ferrites (2237 above) is not correct. Néel's theory, on the contrary, is supported by much experimental evidence. See also 1166, 1170 and 1171 of June, and 3159 of 1949.

538.221 2239

**Magnetic Properties of a Ferrite of Manganese in Weak Fields**—C. Guillaud and A. Barbezat. (*Jour. Rech. Centre. Nat. Rech. Sci.*, no. 11, pp. 83–100; 1950.) Description of the method of measurement of different characteristics, particularly If and Hf losses. Results are presented in tables and graphs.

538.221:537.311.33 2240

**Experimental Study of the Resistance of Some Semiconducting Ferrites**—C. Guillaud and R. Bertrand. (*Jour. Rech. Centre Nat. Rech. Sci.*, no. 11, pp. 73–82; 1950.) Description of method and results of an investigation of Ni, Mn, Mg, and complex ferrites between  $-253^{\circ}$  and  $800^{\circ}\text{C}$ . No exceptional characteristics were observed near the Curie point.

538.221:621.318.22 2241

**Preferred Domain Orientation in Permanent-Magnet Alloys**—M. McCaig. (*Nature* (London), vol. 165, p. 969; June 17, 1950.) Discussion of possible explanations of the improvement in the magnetic properties of alloys such as alcomax due to preferred domain orientation. See also 3453 of 1949 (Hoselitz and McCaig).

546.57:539.23]:537.311.31 2242

**Current/Voltage Characteristic of Very Thin Films of Silver at High Values of Electric Field**—A. Blanc-Lapierre and M. Perrot. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 1641–1643; May 8, 1950.) Measurements were made at ordinary temperatures with field strengths up to 10 kv/cm, on films of equivalent thickness  $<10 \text{ m}\mu$  vaporized on to glass; marked deviations from Ohm's law were observed. The curvature of the characteristic is mainly due to the lacunal structure of the film, heating effects playing only a minor part. Similar characteristics have been observed for semiconductors and for imperfect metal-point contacts.

548.0:535/539 2243

**Optical Properties and the Electro-Optic and Photoelastic Effects in Crystals Expressed in Tensor Form**—W. P. Mason. (*Bell Sys. Tech. Jour.*, vol. 29, pp. 161–188; April, 1950.) The Fresnel ellipsoid is first derived from Maxwell's equations. The electrooptic and photoelastic effects are then expressed in terms of thermodynamic potentials; methods of measuring them by determining the birefringence in various directions are discussed, and constants are listed for various classes of crystal. The application of the photoelectric effect for measuring strains in isotropic media is considered.

620.197 2244

**Tropicalization of Some Materials and Assemblies Used in the Construction of Valves**—

G. Trébuchon. (*Le Vide*, vol. 5, pp. 814–818; May, 1950.) Conclusion of paper noted in 1432 of July.

621.315.33 2245

**Insulated Sleeving and Covered Wires Used on Service Electronic Equipment**—V. G. Hopcroft and G. J. R. Rosevear. (*Proc. IEE* (London), vol. 97, pp. 191–198; May, 1950.) The mechanical and electrical properties desirable on wire-covering material and sleeving are reviewed and standardization is discussed. Summaries are given of interservice specifications. The manufacture and properties of present-day materials are described and their limitations considered. More exacting future requirements are suggested. A table is appended of materials suitable for various temperature and voltage ranges.

621.315.59+551.594.6 2246

**University Research in Physics**—Teegan. (See 2216.)

621.315.59 2247

**Semiconductive Colloidal Suspensions with Non-Linear Properties**—H. E. Hollmann. (*Jour. Appl. Phys.*, vol. 21, pp. 402–413; May, 1950.) The conductivity and dielectric constant of these mixtures (e.g., a fine dielectric powder in oil) depend on the applied electric field; a physical explanation of their properties is given, which is relevant to the general theory of semiconductors. Oscillograms illustrating the nonlinear characteristics of such resistors are shown, and applications are mentioned; the wide range of voltage sensitivity and power dissipation is particularly noted.

621.315.59 2248

**On Electrical Conduction in Semiconductors**—F. Stöckmann. (*Naturwissenschaften*, vol. 37, pp. 85–89 and 105–111; February and March, 1950.) Discussion of conduction processes inside semiconductors, the barrier layer between semiconductors and metals, and photoconductivity.

621.315.59:546.412.64 2249

**Study of the Semiconductor Properties of a Natural Calcium Carbonate**—P. Vidal. (*Ann. Phys.* (Paris), no. 5, pp. 257–309; May and June, 1950.) Experimental and theoretical study of deep-quarried chalk at normal temperatures. Unlike chalk from surface layers, it is relatively pure, has a high resistivity and is a normal semiconductor. Under constant voltage gradient above about 10 v per cm, current intensity decreases with time, tending to a practically constant limit, under which conditions a fall of potential occurs at the cathode. Two methods of determining ion mobility are applied. The rectifying property of the substance and the complex radiation of soft X rays which occurs when a mesh-type cathode is used are discussed.

621.315.612.4 2250

**Deoan Structure and Dielectric Response of Barium Titanate Single Crystals**—B. Matthias and A. von Hippel. (*Phys. Rev.*, vol. 73, pp. 1378–1384; June 1, 1948.)

621.318.22 2251

**The Performance and Stability of Permanent Magnets**—A. J. Tyrrell. (*Jour. Brit. I.R.E.*, vol. 10, pp. 182–191; May, 1950.) The characteristics which contribute to the long-term magnetic stability of a permanent-magnet alloy are considered; present-day anisotropic alloys (e.g., ticonal) embody improvements in all these characteristics. Tests over a period of ten years have proved ticonal magnets to be more stable than materials previously obtainable; they are giving satisfactory service in watt-hour meters, which make high demands on stability. Tables are included giving the performance of numerous British and American commercial alloys.

- 621.357.8:533.5 2252  
**Electrolytic Polishing and Vacuum Technique**—P. Nineuil. (*Le Vide*, vol. 5, pp. 807–813; May, 1950.) Review and discussion of the technique. Its advantages in eliminating surface impurities and reducing gas liberation and "cold" emission, particularly in X-ray tubes, are stressed.
- 666.1.037.5 2253  
**The Physical Aspect of Glass-to-Metal Sealing in the Electronic Valve Industry: Part 1**—G. Reébuchon and J. Kieffer. (*Ann. Radio-élec.*, vol. 5, pp. 125–149; April, 1950.) From a detailed examination of laboratory results together with experience gained in actual manufacture, specifications are drawn up for quality of seal, annealing cycles and finished-work inspection. Physical and chemical aspects of the sealing process are discussed and a critical survey of methods of test is made.
- 669.725.3 2254  
**Beryllium Copper**—J. T. Richards. (*Materials and Methods*, vol. 31, pp. 75–90; April, 1950.) The properties, characteristics, and available forms of the alloy are described, design considerations and methods of treatment discussed, and applications listed.
- 535.37 2255  
**Luminescent Materials [Book Review]**—G. F. J. Garlick. Publisher: Clarendon Press, Oxford, 1949, 254 pp., \$5.50. (*Rev. Sci. Instr.*, vol. 21, pp. 386–387; April, 1950.) "... a book which will give a reader with prior knowledge of solid-state physics an opportunity to become acquainted with some of the main characteristics of luminescence in solid materials together with associated descriptive theory in a minimum of time."
- 666.1.037.5 2256  
**Glass-to-Metal Seals [Book Review]**—J. H. Partridge. Publisher: Society of Glass Technology, Sheffield, 238 pp., 35s. (*Jour. Sci. Instr.*, vol. 27, p. 175; June, 1950.) "A valuable addition to technical literature."
- 517.432.1 2257  
**The Operational Calculus—A Historical Survey of Heaviside's Methods**—(*Electrician*, vol. 144, p. 1751; May 26, 1950.) Report of a paper given by B. van der Pol at one of the IEE Heaviside Centenary Commemorative Meetings, May, 1950. The wide divergence of contemporary estimates of Heaviside's achievements is attributed to the fact that he never gave an exposition of his operational calculus as such; knowledge of it had to be gained by studying the many scattered practical applications in his writings. Modern developments are mentioned, in particular Carson's work in justifying Heaviside's methods by the Laplace transform; Heaviside was aware of this basis to his calculus.
- 517.432.1:621.39 2258  
**Oliver Heaviside's Operational Calculus**—W. Jackson. (*Electrician*, vol. 144, pp. 1663–1667; May 19, 1950.) Heaviside's early experiments on problems in telegraphy, and the development of his mathematical methods, are described.
- 517.564,4:534.26+535.42 2259  
**The Oblate Spheroidal Wave Functions**—A. Leitner and R. D. Spence. (*Jour. Frank. Inst.*, vol. 249, pp. 299–321; April, 1950.) These functions arise in problems such as the diffraction of sound and em waves by circular disks and apertures. Tables published previously by other workers are here extended.
- 517.942.9 2260  
**Separation of Laplace's Equation**—N. Levinson, B. Bogert, and R. M. Redheffer. (*Quart. Appl. Math.*, vol. 7, pp. 241–262; October, 1949.) To use the separation-of-variable method in solving partial differential equations, the surface over which boundary values are specified must coincide with one of the coordinate surfaces, and hence any restriction on the coordinates which can be used is also a restriction on the physical situations to which the method can be applied. The paper is concerned with determining all coordinate systems which are, in this sense, permissible.
- 517.942.932:621.396.611.1 2261  
**Relaxation Oscillations**—J. LaSalle. (*Quart. Appl. Math.*, vol. 7, pp. 1–19; April, 1949.) A class of relaxation oscillations related to the equation  $d^2x/dt^2 + \mu f(x) dx/dt + x = 0$  is studied, in which  $f(x)$  is not necessarily even. Regions  $E(\mu)$  of the phase plane are constructed, each of which contains a single periodic solution. An introductory example deals with the case of an LCR circuit in which  $R$  is nonlinear and non-passive. The use of a series linear resistance to produce mode separation is mentioned.
- 519.283:621.39.001.11 2262  
**Correlation Functions and Communication Applications**—Y. W. Lee and J. B. Wiesner. (*Electronics*, vol. 23, pp. 86–92; June, 1950.) The difference in the forms of the correlation functions for a stationary random process and for a periodic process may be used as a means of separation of the one from the other. The correlation curves are obtained by means of an electronic computer. Among the examples given is the separation of a sinusoidal signal from random noise when an input signal-to-noise ratio of  $-15$  db is changed to an output signal-to-noise ratio of  $+15$  db.
- 681.142 2263  
**High-Speed Digital Calculating Machines**—J. Bennett. (*Distrib. Elec.*, vol. 22, pp. 251–255 and 276–280; March and May, 1950.) A simple account of the historical development and of the operation and capabilities of these machines illustrated by reference to the EDSAC (see also 3458 and 3459 of 1949). The use of acoustic delay lines, special cr tubes and magnetic recording devices for storage has helped to make present-day high speeds possible. Costs are examined and applications, both actual and potential, in research and commerce are mentioned.
- 681.142 2264  
**Linear Equation Solvers**—F. J. Murray. (*Quart. Appl. Math.*, vol. 7, pp. 263–274; October, 1949.) The set of equations to be solved is of the form
- $$\sum_{j=1}^n a_{ij} x_j = b_i$$
- where  $i=1, 2, \dots, n$ . In the type of machine considered, the  $x_i$  unknowns are not driven to their correct values by power supplied from the constant inputs, but reach an equilibrium state corresponding to the solution by a process of adjustment. The operating conditions for such a machine are examined for the case in which the adjusting process is determined by a linear operator with constant coefficients.
- 681.142 2265  
**Pulse Packing in Magnetic Recording Wire**—I. L. Cooter. (*Bur. Stand. Jour. Res.*, vol. 44, pp. 163–174; February, 1950.) An oscillographic method is described for determining the number of magnets per unit length of the recording wire; a powder-pattern method for observing the fields of the individual magnets in a section of wire is also described, with photographs. A comparison was made of the resolution of recorded pulses obtainable with stainless-steel and plated Co-Ni wires. A parameter termed "interference ratio" is defined and used as a criterion of performance. Application is to high-speed digital computers.
- 681.142:533.6 2266  
**Analogue Computer for Flight Simulator**—A. C. Hall. (*Elec. Eng.*, vol. 69, p. 433; May, 1950.) Summary of 1950 AIEE Winter General Meeting paper.
- 681.142:621-526 2267  
**The Electronic Servomechanism Simulator**—C. M. Edwards and E. C. Johnson, Jr. (*Elec. Eng.*, vol. 69, p. 411; May, 1950.) Summary of 1950 AIEE Winter General Meeting paper. The apparatus discussed forms part of a large-scale analogue computer, under development at the MIT, for solving the equations of motion of aircraft. See also above.
- 512.9:501 2268  
**Calcolo Tensoriale e Applicazioni [Book Review]**—B. Finzi and M. Pastori. Publisher: N. Zanichelli, Bologna, 1949, 427 pp., 2,000 lire. (*Quart. Appl. Math.*, vol. 7, p. 352; October, 1949.) The concepts and principal methods of the tensor calculus are presented in order to facilitate its application by mathematicians, physicists, and engineers. The last three chapters deal with applications to (a) the mechanics of deformable continua, (b) electromagnetic theory, and (c) relativity theory.
- 512.972 2269  
**Éléments de Calcul Tensoriel [Book Review]**—A. Lichnerowicz. Publisher: Armand Colin, Paris, 1950, 216 pp., 180 fr. (*Jour. Frank. Inst.*, vol. 249, p. 426; May, 1950.) "The first part comprises an elementary but rigorous presentation of the basic theory both of tensor algebra and tensor analysis. The second part discusses some of the various applications in classical dynamics, electrodynamics, and the theory of relativity."
- 517.53:501 2270  
**Introduction to Complex Variables and Applications [Book Review]**—R. V. Churchill. Publishers: McGraw-Hill, New York and London, 1948, 216 pp., \$3.50. (*Quart. Appl. Math.*, vol. 7, p. 240; July, 1949.) "The book should be of interest to those concerned with the use of function theory in engineering problems and similar applications."
- 517.75:621.3.015.3 2271  
**Transformation Calculus and Electrical Transients [Book Review]**—S. Goldman. Publisher: Constable and Co., London, 1949, 440 pp., 30s. (*Beama Jour.*, vol. 57, p. 150; May, 1950.) The book is written for students, but will also be useful for research workers. Stress is laid upon physical interpretations, and detailed solutions are given of many important examples.
- 517.9:501 2272  
**Introduction to the Differential Equations of Physics [Book Review]**—L. Hopf. Publishers: Dover Publications, New York, 1948, 154 pp., \$1.95. (*Quart. Appl. Math.*, vol. 7, pp. 239–240; July, 1949.) "The exposition is clear and concise, and one might like the treatment to go further than it does. The text should be very useful to seniors and beginning graduate students in physics and chemistry."
- 681.142 2273  
**Giant Brains, or Machines that Think [Book Review]**—E. C. Berkeley. Publishers: J. Wiley, New York, and Chapman and Hall, London, 1949, 270 pp., \$4.00. (*Rev. Sci. Instr.*, vol. 21, pp. 385–386; April, 1950.) A primer dealing with a wide variety of computing mechanisms. "A strong anthropomorphic flavor pervades the book. . . . It contains a useful bibliography, its descriptions of existing machines probably include some facts unknown to any reader, and the discussions of future possibilities, though incomplete, may be suggestive."

## MEASUREMENT AND TEST GEAR

- 531.764.5** 2274  
**Crystal-Controlled Clock**—(Bell Lab. Rec., vol. 28, p. 213; May, 1950.) This instrument was developed as an alternative to the chronometer for use where a portable, accurate standard of time is required. The output of an 1,800-cps crystal-controlled oscillator is used, after frequency division and amplification, to drive a 60-cps clock. Correct time was maintained to within 0.22 sec per day over an 8-day period without temperature control; with temperature control the limit of error could be reduced to 0.09 sec per day.
- 531.764.5** 2275  
**The Performance of the P.T.R. Quartz Clocks and the Yearly Variation of the Length of the Astronomical Day**—A. Scheibe and U. Adelsberger. (*Z. Phys.*, vol. 127, pp. 416-428; March 17, 1950.) Results of time comparisons made from 1934 to 1945, using different types of quartz clock are shown and discussed. Mean daily deviations of the measured time during certain periods are plotted for the different clocks. Observations from 1937, when improved models were brought into operation, confirm the earlier view that the mean daily variation of  $\pm 0.0015$  sec, which has a yearly period, is due to a variation in astronomical standard time, probably corresponding to an actual change in the rotational velocity of the earth. At summer and winter solstices the variation is zero. In summer the mean measured value is up to 0.12 sec in advance of astronomical time.
- 538.71** 2276  
**A Saturated-Core Recording Magnetometer**—D. C. Rose and J. N. Bloom. (*Canad. Jour. Res.*, vol. 28, pp. 153-163; March, 1950.) The magnetometer operates on the saturated-core inductor principle, the sensitive element being a strip of high-permeability alloy. An inverse-feedback system supplies a current to neutralize, within 2 or 3 gammas, the field being measured. This current is measured by a recording milliammeter which gives the field directly.
- 621.317.335.3†+621.317.372** 2277  
**Measuring Dissipation and Dielectric Constants**—C. F. Miller and F. G. Whelan. (*Elec. Eng.*, vol. 69, p. 512; June, 1950.) Summary of A.I.E.E. Winter Meeting paper, giving a description of apparatus used in the susceptance-variation method of measuring dissipation factor due to Hartshorn & Ward (351 of 1937). A constant-speed motor-driven capacitor is used to tune the circuit through resonance. A recording milliammeter operated from a tube voltmeter traces the voltage resonance curves. In 5 runs at 24-hour intervals on each of 35 specimens with dielectric constants in the range 3.0-7.3, the deviation from the mean was  $< 0.05$  in most cases. For the dissipation factor the deviations were generally  $< 2$  per cent.
- 621.317.336:621.315.212** 2278  
**Study of the Impedance Irregularities of Coaxial Cables by Oscillographic Observation of Pulse Echoes**—P. Herreng and J. Ville. (*Ann. Télécommun.*, vol. 3, pp. 317-331; October, 1948.) See 2162 of 1948 and 142 of 1949 (Couanault and Herreng).
- 621.317.336.1** 2279  
**Impedance Measurement**—H. J. Round. (*Wireless Eng.*, vol. 27, pp. 154-158; May, 1950.) A substitution method is described in which the voltages across the unknown impedance and across a known adjustable resistance  $R$  are in turn applied to a cro. A "swamping" series resistance maintains constant current without introducing unduly high errors. Impedances of 2-300  $\Omega$  can be measured over the frequency range 1-150 kc. The vector angle is found by connecting a known capacitance in series with the unknown and obtaining a new value of  $R$ .

- 621.317.351** 2280  
**Various Applications of our Low-Frequency Analyser**—R. Aschen and R. Gaillard. (*TSF Pour Tous*, vol. 26, pp. 170-174; May, 1950.) Oscillograph tracings effected with the instrument described in 2002 of 1948 illustrate the measurement of harmonic distortion, analysis of sawtooth and square-wave signals and their modifications after passing through an amplifier, and analysis of musical sounds.
- 621.317.444** 2281  
**An Electrostatic Fluxmeter of Short Response-Time for Use in Studies of Transient Field-Changes**—D. J. Malan and B. F. J. Schonland. (*Proc. Phys. Soc.*, vol. 63, pp. 402-408; June 1, 1950.) A conducting system of small capacitance is alternately exposed to and screened from the electric field, 1,200 times a second, by a rotating earthed shield and produces an alternating emf of period 0.83 ms. This output is amplified and displayed on a cro screen and can be recorded photographically. An automatic indication of the sense of the field is provided every 7.5 ms. At maximum sensitivity the instrument gives a deflection of 1 cm in a field of 20 v/m with a background noise level of 3 v/m and has a response time of one half-period (0.42 ms).
- 621.396.821:621.317.7.087:551.594.6** 2282  
**On the Recording of the Mean Level of Atmospherics**—F. Carbenay. (*Comp. Rend. Acad. Sci. (Paris)*, vol. 230, pp. 1648-1649; May 8, 1950.) A receiver-recorder has been developed for operation on 27 kc. The receiver has a tuned amplifier stage and a diode detector, and the recorder uses a sensitive highly-damped galvanometer. The mean ordinate of the recorded curve is proportional to the product of the amplitude of the impulses and their repetition rate, i.e., the mean field intensity, while the fluctuations about the mean ordinate are proportional to the individual flux impulses. See also 2902 and 3504 of 1948.
- 621.317.715** 2283  
**An Electromechanism: the Galvanometer**—A. Kaufmann and P. Mériaux. (*Radio Tech. Dig. (Franc)*, vol. 4, pp. 51-61 and 83-89; February and April, 1950.) A mathematical treatment of the theory of the galvanometer, making use of the operational calculus and of electromechanical/electrical analogies. No new properties are shown, but the introduction of complex or operational electromechanical impedances may be useful for the study of new types of galvanometer.
- 621.317.72.088.2** 2284  
**Influence of the Ground on the Calibration and Use of V.H.P. Field-Intensity Meters**—F. M. Greene. (*Bur. Stand. Jour. Res.*, vol. 44, pp. 123-130; February, 1950.) The input impedance of receiving antennas varies with height and changing ground conditions. An approximate method is described for calculating this impedance in the case of horizontal dipole antennas over earth having finite values of dielectric constant and conductivity. The effect of changes in ground conditions and antenna terminating impedance on the error of the field-intensity meter is calculated as a function of antenna height. Measured error values are in reasonably good agreement with the theoretical values.
- 621.317.725.029.45/c5** 2285  
**A Millivoltmeter for the Frequency Range From 1,000 to  $30 \times 10^6$  c/s**—H. J. Lindenhovius, G. Arbelet, and J. C. van der Breggen. (*Philips Tech. Rev.*, vol. 11, pp. 206-214; January, 1950.) Six inductance-compensated RC amplifying stages between the input stage and a moving-coil meter with series crystal rectifier give an over-all amplification of about 1,500 throughout the frequency range. Compensation for mains fluctuation and tube aging is provided.

vided. Voltage ranges are 1 mv, 10 mv, and thence by successive factors of  $\sqrt{10}$  to 1 kv, the range being selected by adjustment of a capacitive attenuator.

- 621.317.73** 2286  
**A Direct-Reading Instrument for Measuring Unbalanced Impedances at Decimetric Wavelengths**—W. H. Ward. (*Proc. IEE (London)*, vol. 97, pp. 199-205; May, 1950.) The instrument is based on the reactance-variation method, the circuit elements being lengths of concentric line. Direct voltage measurement is eliminated and replaced by the reading of a piston attenuator. By enclosing the oscillator in the moving part of the attenuator, the use of flexible hf cable is avoided. The mechanical drive may, if required, be transmitted from a distance. The instrument measures capacitance from -10 pF to +10 pF and resistance from 10 to 10,000  $\Omega$  at frequencies up to at least 1,000 Mc, with an accuracy within about 1 per cent. Both scales are direct-reading.
- 621.317.734:621.319.4** 2287  
**A Direct Reading Instrument for the Measurement of the Series-Resistance of Capacitors**—F. Gutmann. (*Jour. Sci. Instr.*, vol. 27, pp. 169-170; June, 1950.) A thermocouple ammeter measures the rf tank current of a 1.6-Mc oscillator, whose output is adjusted to produce full-scale deflection. The capacitor under test is then inserted in series with the tank capacitor. The resultant decrease of tank current is a measure of the series resistance of the tested capacitor. Capacitors from 0.01  $\mu$ F upwards may be tested.
- 621.317.755:621.385.012** 2288  
**Curve Generator for Electron Tubes**—(*Tech. Bull. Nat. Bur. Stand.*, vol. 34, p. 62; May, 1950.) Equipment developed by M. L. Kuder.  $I_a/V_a$  or  $I_a/V_g$  families of characteristics are displayed simultaneously on a cro to an over-all accuracy within  $\pm 5$  per cent on voltage and current readings. Measurement is by a standard co-ordinate rectangle. Marker dots indicate the load-line for the conditions selected. Special circuits are included in the equipment for observation of the step-function signal at the grid of the tube under test and for identifying the curves obtained with positive grid bias.
- 621.317.772** 2289  
**Direct-Reading Phasemeter**—L. H. O'Neill and J. L. West. (*Rev. Sci. Instr.*, vol. 21, pp. 271-273; April, 1950.) The amplitude of the sum or difference of two sinusoidal voltages of equal amplitude depends on the phase difference between them; this is the basis of the simple phase meter here described. The method of adjusting the individual amplitudes and mixing the two voltages, and of determining the sign of the phase difference, are discussed. The over-all accuracy is within  $\pm 1^\circ$  over the frequency range 7 cps to 100 kc.
- 621.317.79.001.4:621.397.6** 2290  
**Television Monitors**—J. E. B. Jacob. (*Wireless World*, vol. 56, pp. 206-210; June, 1950.) Equipment of higher quality than that of the average home set is required. The special features that are necessary in the design of the magnetic lens, focus-correction system, synchronizing-signal separator and black-level control circuit are described.
- 621.385.18.001.4** 2291  
**Measurement of Tube Voltage Drop in Hot-Cathode Gas Tubes**—E. K. Smith. (*Elec. Eng.*, vol. 69, pp. 419-422; May, 1950.) 1950 AIEE Winter General Meeting paper. Increase of the voltage drop is one of the commonest preliminary indications of failure in gas-filled tubes. Various methods are described for observing this drop while the tubes are in use, typical results are charted, and their significance discussed.

621.396.615:621.317.7.001.4 2292

**VOR Signal Generator**—J. H. Battison. (*Tele-Tech*, vol. 9, pp. 37–39, 57; May, 1950.) Operation of the VOR is based on the phase difference between two 30-cps signals, one with a rotating antenna pattern and the other an omnidirectional reference. The apparatus described generates such signals at a strength sufficient for testing the receiving equipment of aircraft in a nearby field or hangar. The generator can also be used to test runway localizers in which the reference signal is in phase with the variable signal on one side of the runway and 180° out of phase on the other.

621.396.615.029.42 2293

**A Low-Frequency Sinusoidal-Voltage Source**—L. A. Rosenthal. (*Rev. Sci. Instr.*, vol. 21, pp. 302–303; April, 1950.) Frequencies extending from 20 cps nearly to zero can be obtained by a method in which a variable-frequency audio oscillator is heterodyned with a stable frequency derived from ac mains. A mathematical explanation of the mixing operation is given. Application to the testing of a voltage regulator is mentioned.

621.396.615.14.029.63 2294

**U. H. F. Sweep-Frequency Oscillator**—J. E. Ebert and H. A. Finke. (*Electronics*, vol. 23, pp. 79–81; June, 1950.) A variable capacitor, with vane rotated at the high-impedance end of a cavity resonator by a synchronous motor, provides a maximum sweep of 30 Mc on the fundamental frequency of an enclosed coaxial-line type of oscillator covering a range of 470–890 Mc. The sweep can be varied by altering the capacitor plate spacing. The output voltage is continuously variable from 2 v across 50  $\Omega$  to a value 90 db below this. Details of the equipment are given, together with performance curves and a dimensioned sketch of the oscillator.

621.396.822:621.316.8 2295

**Resistor Noise**—(*Radio Tech. Dig.* (Frang.), vol. 4, pp. 91–107; April, 1950.) French version of 1465 of July (Paolini and Canegallo).

621.397.645:621.316.761.2 2296

**How to Adjust Frequency Response in Video Amplifiers for TV**—J. H. Roe. (*Broadcast News*, no. 58, pp. 54–65; March and April, 1950.) Methods of adjusting video amplifier circuits to obtain satisfactory frequency response characteristics, and methods of measuring their performance, are described with practical details of the necessary equipment. Typical results are illustrated by numerous oscillograms showing the effects of correct alignment and of misadjustment of a particular circuit commonly used for hf and lf compensation.

621.317 2297

**Electrical Measurements in Theory and Application [Book Review]**—A. W. Smith. Publisher: McGraw-Hill, London, 1948, 371 pp., 25s. 6d. (*Beama Jour.*, vol. 57, pp. 150–151; May, 1950.) Though mainly addressed to students, the book is recommended to laboratory and test-room workers in general. New methods of measuring capacitance and inductance are included in this edition.

621.317.755 2298

**Cathode-Ray Tube Traces [Book Review]**—H. Moss. Publisher: Electronic Engineering, London, 66 pp., 10s. 6d. (*Jour. Sci. Instr.*, vol. 27, p. 175; June, 1950.) "In putting the main emphasis on the geometrical theory of cathode-ray-tube patterns, Dr. Moss has produced a monograph which will be valuable both to the advanced student and to the qualified engineer."

621.396.001.4 2299

**Radio Servicing Equipment [Book Review]**

—E. J. G. Lewis. Publisher: Chapman and Hall, London, 371 pp., 25s. (*Jour. Sci. Instr.*, vol. 27, p. 175; June, 1950.) "Describes in plain and simple language the construction and operation of most of the types of test and measuring equipment used by the radio service engineer."

### OTHER APPLICATIONS OF RADIO AND ELECTRONICS

529.781 2300

**Carrier Frequency for Remote Timing Control**—J. L. Wagner and W. Webb. (*Elec. Mfg.*, vol. 43, pp. 104–107, 198; April, 1949.) The requirements for a timing-control system operating on power distribution lines are enumerated. A detailed description is given of commercial equipment in which the master clock operates keying relays in the transmitter, providing three separate channel frequencies in the range 2–6 kc, one for time correction of slave clocks and the others for applying pulses at 1-minute intervals to attendance recorders, for ringing bells, etc.

534.321.9:615 2301

**Ultrasonics in Therapy**—P. Hémardinger. (*Rev. Belge Électronique*, vol. 1, pp. 14–21; February, 1950.) A review of the technique, and description of typical generating apparatus.

535.322.1:539.165 2302

**On an Improvement in Lens-Type  $\beta$ -Ray Spectrographs**—P. Grivet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 1652–1654; May 8, 1950.)

539.16.08 2303

**Radiation Counters**—Please note that, in this and subsequent issues, the UDC number used will be 621.387† and subdivisions, instead of 539.16.08 used hitherto.

621.317.083.7 2304

**Present-Day Telemetry Technique**—G. Swoboda. (*Elektrotech. u. Maschinenb.*, vol. 67, pp. 129–136; May, 1950.) Review of current methods and description of a pulse system and of one using rotary selector switches.

621.317.087.9:621.317.083.7 2305

**The Metrotype System of Digital Recording and Telemetry**—G. E. Foster. (*Elec. Eng.*, vol. 69, pp. 427–430; May, 1950.) 1950 AIEE Winter General Meeting paper. The system enables electrical quantities to be automatically measured and recorded in numerals, at predetermined intervals. Readings may be transmitted to any distance, a single channel of telegraph quality being adequate. The processes of measuring, transmitting, and printing the readings are described.

621.317.39 2306

**Electrical Methods of Measuring Mechanical Quantities**—(*Engineering* (London), vol. 169, pp. 483–484; April 28, 1950.) Report of a paper given at the IEE by F. J. Woodcock.

621.365.55†:674.23 2307

**High-Frequency Generators: Application in the Wood Industry**—M. Diopère. (*Rev. Belge Électronique*, vol. 1, pp. 22–28; February, 1950.) Description of different manufacturing processes, e.g., bending, sticking, etc., in which hf heating is applied to materials under pressure.

621.38.001.8:543/545 2308

**Electronic Instrumentation for Chemical Laboratories**—F. Gutmann. (*Jour. Brit. I.R.E.*, vol. 10, pp. 194–212; May, 1950.) Reprint. See 710 of April.

621.384.611.2† 2309

**The Proton Synchrotron**—E. J. Lofgren. (*Science*, vol. 111, p. 295–300; March 24, 1950.) Brief survey of development of particle ac-

celerators and description of the proton synchrotron now under construction at Berkeley, which will be ultimately capable of producing protons of about  $6 \times 10^3$  Mev.

621.384.612.1† 2310

**Omegatron—a Miniature Cyclotron**—(*Electronics*, vol. 23, pp. 122, 124; June, 1950.) The omegatron operates on the same fundamental principle as the cyclotron and can be used as a mass analyzer. In addition to the constant uniform magnetic field a rf electric field is applied at right angles. When the rf equals the cyclotron frequency the ions spiral outwards and can be collected. Measurement of the frequency corresponding to maximum ion current and a measurement of the magnetic field enable the mass of the particles to be determined.

621.385:621.318.572 2311

**An Electron-Beam Decimal-Counter Tube**—Hollway. (See 2374.)

621.385.833 2312

**On Magnetic Cylindrical Lenses with Image-Distortion Correction**—H. Hintenberger. (*Z. Naturf.*, vol. 3a, pp. 125–127; February, 1948.) The conditions for exact focusing are determined and also for first- and second-order approximations.

621.385.833 2313

**The Calculation of Magnetic-Lens Fields by Relaxation Methods**—M. B. Hesse. (*Proc. Phys. Soc.*, vol. 63, pp. 386–401; June 1, 1950.) "The field and lens constants are calculated for two typical magnetic lenses as used in electron microscopes, using the relaxation method developed by Southwell for the solution of potential problems."

621.385.833 2314

**The Independent Electrostatic Lens under 'Transgaussian' Conditions**—E. Regenstein. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 1650–1652; May 8, 1950.) A study of the paths of rays termed transgaussian which are initially parallel to the axis but for which  $r_0/z_0$  is negligible.

621.385.833:621.396.615.141.2 2315

**Electron-Optical Mapping of the Space-Charge Field in a Magnetron**—(*Tech. Bull. Nat. Bur. Stand.*, vol. 34, pp. 57–59; May, 1950.) Method devised by D. L. Reverdin. Space-charge distribution is related to the distorting effect of the electric field in an elementary steady-stage magnetron on the shadow pattern formed by two wire screens in the path of a focused non-disturbing axial electron beam. The method may be extended to oscillating magnetrons and to the indication of noise. Results differ widely from predictions based on theory.

621.395.625.3.001.8 2316

**Various Applications of Magnetic Recording**—P. Hémardinger. (*Radio Franç.* no. 6 pp. 8–16; June, 1950.) Description of the principles and basic circuits of magnetic recorders for secret transmission phototelegraphy, time measurement, etc.

629.13.014.57 2317

**The Electronic Automatic Pilot**—M. Nerinckx. (*Rev. Belge Électronique*, vol. 1, pp. 5–13; March and April, 1950.) Description of the Sperry automatic pilot Type A-12.

621.38.001.8 2318

**Industrial Electronics [Book Review]**—A. W. Kramer. Publisher: Pitman Publishing Co., New York, 1949, 311 pp., \$6.00. (*Electronics*, vol. 23, pp. 136, 140; June, 1950.) "... intended for readers who have a good knowledge of general physics and engineering but who have had very little training or experience in electronics."

- 621.387  
Counting Tubes [Book Review]—S. C. Curran and J. D. Craggs. Publisher: Butterworths Scientific Publications, London, 238 pp., 35s. (*Jour. Sci. Instr.*, vol. 27, p. 175; June, 1950.)

## PROPAGATION OF WAVES

- 538.56:535.42 2320  
On the Diffraction of an Electromagnetic Wave through a Plane Screen—J. W. Miles. (*Jour. Appl. Phys.*, vol. 21, p. 468; May, 1950.) Corrections to paper abstracted in 202 of February.

- 621.396.11 2321  
The First Ionospheric-Storm-Warning Service 1941-45—J. S. Kojan and G. A. Isted. (*Marconi Rev.*, vol. 13, pp. 53-71; 1950.) The development of this service at Baddow is traced from its origin during the War, and an account is given of the factors upon which the forecasts are based. The efficiency of the storm warnings is reviewed and new developments are discussed in connection with future storm-warning services; an "idealized" warning system might include: solar-corona observations; solar-noise observations; monitoring of radio circuits having great-circle paths in (or near) the auroral zones; geomagnetic measurements; sunspot observations.

- 621.396.11:551.510.535 2322  
The Air Force Interest in Sporadic E Ionization—Gerson. (See 2212.)

- 621.396.81 2323  
Reception [at Madrid] of [s.w.] Transmissions from Great Britain, the Vatican and U.S.A.—R. G. Sacasa. (*Rev. Telecomunicación* (Madrid), vol. 4, pp. 10-16; December, 1948.) Graphs show receiver output power variations during March, 1948, at various times of the day for transmissions in the range from 16.85 m to 50.26 m. Average values are compared with those for signals from Spanish and Australian stations.

- 621.396.81 2324  
Washington and Madrid [radio] Propagation Forecasts—R. G. Sacasa. (*Rev. Telecomunicación* (Madrid), vol. 4, pp. 20-29; June, 1949.) A comparison is made between NBS forecast charts and the author's, based on experimental observations, for reception of wavelengths in the range 13-31 m. Agreement for January and February, 1949, was better than for 1942, but there were still appreciable discrepancies.

- 621.396.81 2325  
Back-Scatter Observations by the Central Radio Propagation Laboratory—August 1947 to March 1948—W. L. Hartsfield, S. M. Ostrow, and R. Silberstein. (*Bur. Stand. Jour. Res.*, vol. 44, pp. 199-214; February, 1950.) A high-power pulsed transmitter, the circuit diagram of which is given, and westwardly beamed antenna system were used, transmissions being made from Sterling, Virginia, at a frequency of 13.66 Mc. Back-scatter observations are presented graphically and photographs are shown of different types of echo pattern received. Back-scatter from both the ground and the E region was identified, ground scatter usually predominating. Comparisons of results are made with transponder signals and skip distance maps drawn from concurrent ionospheric data. Curves of calculated transmission and echo delays are also given.

- 621.396.81 2326  
Effect of a Sudden Ionospheric Disturbance on Long Radio Waves Reflected Obliquely from the Ionosphere—K. Weekes. (*Nature* (London), vol. 165, pp. 935-936; June 10, 1950.) The behavior of radio waves reflected at oblique or at steep incidence during periods of

sudden ionospheric disturbance is outlined. Signal-strength measurements on the Danish transmissions for the Decca navigation system suggest that for frequencies of about 100 kc, transmitted over about 900 km, the reflected wave amplitude is increased during a sudden ionospheric disturbance, corresponding to an effective reflection coefficient of 0.25, which is about five times the normal value. When waves of the same frequency are reflected at steep incidence, a decrease in amplitude is observed. In either case, the associated phase change corresponds to a decrease in the height of reflection.

On frequencies above 2 Mc a decrease in amplitude occurs both for vertical and oblique incidence, whereas frequencies in the region of 16 kc show little amplitude change at steep incidence but an increase at oblique incidence. The above results confirm and extend the recent observations of Smith-Rose (1228 of June).

- 621.396.812 2327  
A Survey of Ionospheric Cross-Modulation—L. G. H. Huxley and J. A. Ratcliffe. (*Proc. IEE* (London) vol. 97, p. 165; May, 1950.) Discussion on 211 of February.

- 621.396.82 2328  
Contribution to the Study of Gyromagnetic Frequency by means of Gyro-Interaction of Radio Waves in the Ionosphere—M. Cutolo. (*Nuovo Cim.*, vol. 5, pp. 475-488; October 1, 1948.) A full account of experiments carried out in 1947 and briefly reported in 2055 of 1948. The results confirm the previous observation that there is a regular rise in the value of the resonance frequency during the course of the night.

- 621.396.812 2329  
Low-Level Atmospheric Ducts—R. F. Jones. (*Nature* (London), vol. 165, p. 971; June 17, 1950.) The variations in signal strength over paths in the Cardigan Bay area in conditions of strong wind can be related to the origin and subsequent track of the air over that area at the time. In particular, the drying effect on the lowest layers due to recent travel over land is important. See also 2033 of 1949.

- 551.510.535:621.396.11 2330  
Short-Wave Radio and the Ionosphere [Book Review]—Bennington. (See 2218.)

## RECEPTION

- 621.396.621 2331  
The Transmission-Line Discriminator—P. Magne. (*Ann. Radioélec.*, vol. 5, pp. 89-93; April, 1950.) The principle is described by which discrimination is effected with an open-ended and a closed coaxial line across a common fm source. Calculation is made of the length of line required and of the amount of third harmonic. A harmonic-filter circuit including two tubes Type 6AK5 is described for a discriminator for use in a 20-Mc band centered at 105 Mc. The response curve is flat over the whole band and distortion is of the order of 0.1 per cent.

- 621.396.621.5 2332  
A Crystal Tetrode used as a Frequency Changer—R. W. Haegele. (*Onde Élec.*, vol. 30, pp. 239-241; May, 1950.) Preliminary experimental study showing the conversion characteristics as functions of the collector current, local oscillator voltage, etc. The crystals were of the type normally used in Ge diodes Type IN34, the points of contact of the three electrodes at the crystal surface forming an equilateral triangle of side about 0.05 mm. Signal frequencies up to at least 200 Mc may be used; there is little interaction between the input circuits; conversion gain is high. The maximum if appears to be sensibly equal to the maximum

frequency of amplifying crystal triodes, so that the input signals may have much higher frequencies.

- 621.396.821:621.317.7.087:551.594.6 2333  
On the Recording of the Mean Level of Atmospherics—Carbenay. (See 2282.)

- 621.396.822 2334  
Thermal Noise at High Frequencies—A. van der Ziel. (*Jour. Appl. Phys.*, vol. 21, pp. 399-401; May, 1950.) "Spence's discussion of thermal noise from the point of view of the electron theory leads to a result which differs from Nyquist's original formula at frequencies  $\nu$  such that  $h\nu/kT \geq 1$ . This discrepancy is due to the assumptions which Spence made in his analysis."

- 621.396.822:621.396.619.13 2335  
The Spectrum of Frequency-Modulated Waves after Reception in Random Noise: Parts 1 & 2—D. Middleton. (*Quart. Appl. Math.*, vol. 7, pp. 129-174; July, 1949; and vol. 8, pp. 59-80; April, 1950.) Mathematical analysis. The wave emerging from the if amplifier and entering the limiter is assumed to be of the narrow-band type, the limiter providing an output voltage proportional to the instantaneous amplitude of the wave for amplitudes less than a certain threshold value, and a constant output for larger amplitudes. The discriminator is "ideal" and provides an output voltage directly proportional to the instantaneous difference in frequency between the input wave and the central if.

The spectrum of the signal and of the noise emerging from the discriminator is analyzed generally. The results of the analysis for no limiting and extreme limiting, for different input signal-to-noise ratios and for carrier mistuning, are presented graphically. Part 2 extends the theory given above to the general case of an arbitrary degree of limiting, the postulated conditions being the same as before. The spread of the spectrum of the noise due to the clipping produced by the limiter is evaluated. The results for various degrees of limiting are presented for the two special cases of no carrier (noise alone) and strong carrier.

- 621.396.621+621.396.61 2336  
Radio Receivers and Transmitters [Book Review]—S. W. Amos and F. W. Kellaway. Publisher: Chapman and Hall, London, 2nd edn. revised 1948, 356 pp., 25s. (*Beama Jour.*, vol. 57, pp. 117-118; April, 1950.) Deals mainly with receivers, only one of the ten chapters being devoted to transmitters. "... this book contains much useful information, and where proofs are lacking references are indicated."

- 621.396.621+621.396.645]:621.385 2337  
Electronic Valves: Books II-IV [Book Review]—(See 2388.)

## STATIONS AND COMMUNICATION SYSTEMS

- 621.39.001.11 2338  
On the Maximum Transmission Capacity of a Channel in the Presence of Noise—J. Laplume. (*Onde Élec.*, vol. 30, pp. 235-238; May, 1950.) Paper presented at the Congrès d'Électronique et de Radioélectricité, January, 1950. Random noise is the only essential factor limiting the capacity of a channel. A formula is derived showing the number of signals which may be transmitted in a given time interval. In the ideal case this number is an exponential function of the bandwidth. In the case of AM or FM the number is proportional to the pass-band. With PCM the theoretical limit may be reached. Quantization of a continuous message is possible only if the channel bandwidth is greater than that necessary for ssb AM transmission, the bandwidth being increased at the expense of transmission power. See also 1649 of 1949 (Shannon) and 2319 of 1949 (Tuller).

- 621.39.001.11:519.283 2339  
Correlation Functions and Communication Applications—Lee and Wiesner (See 2262.)

621.394/.396 2340  
Telecommunications and Heaviside—(Electrician, vol. 144, p. 1752; May 26, 1950.) Report of a paper "Fifty Years' Development in Telephone and Telegraph Transmission in Relation to the Work of Heaviside," by W. G. Radley, given at one of the IEE Heaviside Centenary Commemorative Meetings, May, 1950. Heaviside in 1876 first realized the need for an inductance term and, subsequently, for a leakage-conductance term in the basic equations for line transmission; the use of loading coils as standard practice followed from his theory. Modern improvements of these coils are mentioned. Heaviside's work on em waves in space is considered and an account is given of systems developed to overcome the effects of ionospheric irregularities on long-distance radio communication.

621.395 2341  
The Telephone and the Laboratory: Part 1—H. M. Trainor. (Trans. S. Afr. Inst. Elec. Eng., vol. 40, pp. 251-274; November 11, 1949. Discussion, pp. 275-280.) The organization of telecommunications research in the Bell System, U.S.A., the British Post Office, and the Australian Post Office is reviewed, and an account is given of work done in South Africa. This has consisted largely in the quality testing of equipment and the investigation of problems arising from the special conditions of use existing in South Africa. Laboratory facilities of the Post Office and other services are described and various laboratory and field investigations are discussed. Part 2: 2342 below.

621.395 2342  
The Telephone and the Laboratory: Part 2—H. M. Trainor. (Trans. S. Afr. Inst. Elec. Eng., vol. 41, pp. 57-79; March, 1950. Discussion, pp. 80-86.) Relevant theories of speech, hearing, and noise are reviewed and a detailed discussion is given of the transmission aspects of the telephone instrument, which is a source of distortion in a telephone system. The development of the modern instrument has been based on a detailed knowledge of speech characteristics and the various noise components affecting reproduction. The carbon-type transmitter and the importance of improved magnetic materials in reduction of frequency distortion in receivers are discussed. Side tone can be reduced by modification of the matching induction-coil circuit. Tests of sensitivity, frequency response, nonlinear and amplitude distortion and articulation, are important measures of the efficiency of a system. The over-all effective transmission or repetition rating, as determined in practice, is the final criterion of efficiency. Part 1: 2341 above.

621.396.1 2343  
Technical Repercussions of the Copenhagen Plan—J. Garcin. (Toute la Radio, vol. 17, pp. 152-154; May, 1950.) Comment on the practical effects of the redistribution of wavelengths for French broadcasting, and suggested receiver adjustments to minimize interference.

621.396.619.16 2344  
Pulse-Time and Pulse-Width Modulation—P. Bréant. (Ann. Télécommun., vol. 3, pp. 309-316; October, 1948.) Theoretical study made in 1946. The systems are defined and a rapid method for determining the modulation spectrum is described. In the case of PTM, demodulation cannot be effected with a simple low-pass filter. In PWM this demodulation method is effective only if (a) modulation depth is small; (b) the sides of the pulses displaced by the modulation are steep; and (c) only one side of the pulse is displaced. Demodulation by

simple filtering is thus of little practical use for PM signals.

621.396.619.16 2345  
Narrow-Band Pulse Communication—T. Roddam. (Wireless World, vol. 56, pp. 202-205; June, 1950.) Short square-wave amplitude-modulated pulses are converted to Gaussian pulses; the most efficient method is to use an amplifier with several "maximal flatness" stages {3013 of 1941 (London)} and one plain RC stage. After the Gaussian pulses have traveled along telephone cables provided with suitable intermediate repeaters, the shape is preserved but there is overlapping of the pulse signals. The overlapping increases as the bandwidth is decreased, causing cross-modulation, but a corrector circuit can be used to eliminate this effect.

621.396.65 2346  
Carrier Power Requirements for Long-Distance Communication by Microwaves—A. G. Clavier. (Elec. Commun., vol. 27, pp. 39-47; March, 1950.) A general expression is given for the path attenuation of a radio link in terms of wavelength, distance, and effective areas of transmitting and receiving antennas; abacs are presented for finding values of the free-space attenuation between (a) nondirectional and (b) parabolic antennas. Conditions for PCM and FM transmission are compared; the rf power required for a PCM system is 17 db less than for a corresponding FM system, suitable allowances being made for fading and for receiver noise, but under conditions of deep fading the FM system is the better.

621.396.65 2347  
A Simple Microwave Relay Communication System—M. G. Staton. (Tele-Tech, vol. 9, pp. 40-42, 44; April, 1950.) This equipment operates in the 952-960-Mc band and uses conventional tubes and circuits. The carrier frequency is derived by repeated multiplication from a p.h.m. crystal-controlled oscillator operating at about 1 Mc, and the transmitter output is  $\leq 3$  W. The receiver is a superheterodyne with self-contained power supply. Dipole antennas with 42-in. parabolic reflectors are used for both transmitter and receiver. Unattended repeater stations are provided with automatic monitoring arrangements.

621.396.65 2348  
V. H. F. Links at Manila Airport—E. J. Rudisuhle and P. B. Patton. (Electronics, vol. 23, pp. 82-85; June, 1950.) Line-of-sight FM links in the 160-Mc band provide over 150 speech, telegraphy, printer, and control circuits between the control station and remote transmitting and receiving stations. Details are given of some of the special circuits used.

621.396.823:621.395.44 2349  
On Radiation from Overhead Transmission Lines—M. Janssen. (Proc. IEE (London), vol. 97, pp. 166-178; May, 1950.) An analysis is made of the fields of single and multiple lines over imperfectly conducting earth. Field strengths measured along profiles transverse to a 3-phase line in a valley agreed quite well with theoretical values, for different connections of the lines. Single phase-to-earth connection in power-line carrier systems is undesirable if interference is to be minimized. It is suggested that horizontal antennas may be useful for low-power broadcasting transmission in valleys.

621.396.931 2350  
The Introduction in Switzerland of the Public Telephone Service to Vehicles—H. Kappeler. (Bull. Schweiz. Elektrotech. Ver., vol. 40, pp. 433-439; July 9, 1949. In French, with German summary.) By the service introduced at Zürich in June, 1949, 2-way telephone connection via a radio link may be made with any suitable equipped vehicle plying within

about 10 km of a fixed transmitting/receiving station. Details are given of the automatic calling and selection system and of the radio-telephone equipment. Phase modulation is used, with carrier frequencies between 31 and 41 Mc.

621.396.932:621.396.5.029.62 2351  
Thames Radio Service—(Wireless World, vol. 56, pp. 215-216; June, 1950.) For an abstract of another account see 999 of May (Neale and Burr).

## SUBSIDIARY APPARATUS

621-526:621.396.645.37 2352  
Feedback Amplifiers and Servo Systems—Ward. (See 2175.)

621-526:621.398 2353  
Synco Units for Remote Indication and Control—S. N. Mead and G. E. Day. (Elec. Mfg., vol. 43, pp. 74-79, 204; April, 1949.) Operating characteristics are given of many types of ac self-synchronous transmitter and indicator units commercially available, together with design details useful in selecting equipment for particular practical applications.

621.3.077.2/.3:621.3.016.35 2354  
On the Application of Stability Criteria to an Amplidyne Control Circuit—F. van Geerttruyden. (HF (Brussels), no. 6, pp. 169-176; 1950.) The amplidyne, a dynamoelectric amplifier, plays a part in control circuits very similar to that of an electronic amplifier. The mathematical study of such amplifiers, particularly with reference to stability, has, been extended to the amplidyne. Two stability criteria are examined: that of Leonhard, which is mathematically tractable, and the generalized Nyquist criterion, which is particularly suitable for use in the experimental study of a control circuit. The formulation of the equation for a control circuit is simple in operational notation. Transients superposed on the steady state are specially studied. An account is given of a complete numerical investigation, both theoretical and experimental, of the stability in two types of alternator-voltage regulation. In one case the amplidyne supplies the alternator excitation directly and in the other case feeds a separate exciter.

621.316.726:621.314.3† 2355  
A Magnetic-Amplifier Frequency Control—L. J. Johnson and H. G. Schafer. (Elec. Eng., vol. 69, p. 445; May, 1950.) Summary of 1950 AIEE Winter General Meeting paper. A circuit is described for stabilizing the 60-cps output frequency of a small motor-driven alternator. A magnetic amplifier is controlled by the output of a pair of balanced tuned circuits fed from a pilot generator attached to the dynamotor shaft, and its output in turn regulates the field of the dynamotor.

Temperature stability is obtained by the use of permalloy cores in the tuned circuits. Frequency regulation of the order of 1 part in 5,000 at normal temperatures is claimed.

621.396.682 2356  
Design, Construction and Test of a Stabilized Anode Voltage [supply unit]—L. Chrétien. (TSF Pour Tous, vol. 26, pp. 177-180; May, 1950.) Details and performance of equipment using the circuit noted in 1532 of July.

621.398 2357  
Radio Synchro-Motor [Selsyn]—In 1258 of June please change J. R. Duthil to J. R. Dutilh.

## TELEVISION AND PHOTOTELEGRAPHY

621.397.6 2358  
Line-by-Line Black-Level Control of Television Signals—N. N. Parker Smith. (Marconi Rev., vol. 13, pp. 81-85; 1950.) The gain of a receiver or the carrier level of a transmitter is controlled by a voltage derived during part of

the 5- $\mu$ s black-level period at the commencement of each line. This control voltage, proportional in amplitude to the black-level signal, remains steady over the line period of approximately 100  $\mu$ s, and yet follows any changes which occur between successive lines. The description of the control arrangement is illustrated by circuit and waveform diagrams.

**621.397.6** 2359  
Simplified Television for Industry—R. C. Webb and J. M. Morgan. (*Electronics*, vol. 23, pp. 70-73; June, 1950.) Equipment using the vidicon camera tube (2040 of September).

**621.317.79.001.4:621.397.6** 2360  
Television Monitors—Jacob. (See 2290.)

**621.397.611.21.002.2** 2361  
Producing the 5820 Image Orthicon—R. B. Janes, R. E. Johnson, and R. R. Handel. (*Electronics*, vol. 23, pp. 93-95; June, 1950.) Short illustrated account of the special techniques used.

**621.397.62** 2362  
Magnetic Deflector Coils—(*Radio Tech. Dig.* (Franc), vol. 4, pp. 131-156; June, 1950.) An article based primarily on 2043 of September (Cocking), with added bibliography.

**621.397.621.2** 2363  
R.C.A. Color Kinescope Demonstrated—(*Tele-Tech*, vol. 9, pp. 20-21, 63; May, 1950.) In the two laboratory-produced tubes described the screen is composed of 351,000 color-emitting dots arranged in triangular groups of 3, corresponding to the three primary colors of the picture. An apertured mask is interposed between the electron beam source and the screen so that one color only is emitted by each group as it is scanned. (a) Three-gun tube. Each gun is operated in the time sequence corresponding to the sampling process at the transmitter. Each gun excites one color only on the tube face. (b) Single-gun tube. Two sets of coils fed with quadrature currents at the sampling frequency generate a rotating field producing circular deflection of the beam. The phase of the field determines which color is excited on the tube face.

**621.397.645:621.316.761.2** 2364  
How to Adjust Frequency Response in Video Amplifiers for TV—Roe. (See 2296.)

**621.397.743:621.392.5:621.315.212** 2365  
Equalization of Coaxial Lines—Gould. (See 2152.)

**621.397.82:621.392.52** 2366  
The Latest Techniques for the Elimination of Ham TVI—Rand. (See 2156.)

**621.397.5** 2367  
Television Explained [Book Review]—W. E. Miller. Publisher: The Trader Publishing Co., London, 3rd edn., 104 pp., 5s. (*Jour. Brit. I.R.E.*, vol. 10, p. vi; May, 1950.) A number of sections have been revised and additional information is included. "The treatment is non-mathematical.... The student, the radio service engineer, and the amateur will find this book an excellent investment...."

## TRANSMISSION

**621.394.61** 2368  
Low-Frequency Radiotelegraph Transmitter for Arctic Use—H. P. Miller, Jr. (*Elec. Commun.*, vol. 27, pp. 11-20; March, 1950.) A conventional 10-kw equipment for operation in the frequency range 80-200 kc or, in later models, 200-300 kc. Single-frequency operation with on-off keying is normally used, but provision is also made for modulated cw operation. Arctic conditions necessitate air cooling of the F.892R final amplifier tube and preheating of the Hg-vapor rectifiers. Unit construction is

employed, to facilitate the transportation of the equipment by air.

**621.396.61:621.396.97** 2369  
Contribution to the Study of High-Efficiency Broadcasting Transmitters—J. Polonsky. (*Ann. Radioélec.*, vol. 5, pp. 109-124; April, 1950.) The simplified theory of the grid-modulated amplifier is criticized and a more rigorous theory put forward. The disadvantages of the Doherty-Terman circuit (4020 of 1936 and 4300 of 1938) are practically eliminated in the experimental SIF transmitter at Boutigny by (a) locating the impedance inverter at the rf exciter input; (b) mounting the two modulator tubes in a cathodyne connection; and (c) incorporating an over-all degenerative and a local regenerative feedback loop. Useful carrier power of this transmitter is 18 kw; plate efficiency of the final stage 73 per cent; over-all efficiency 43 per cent. Other characteristics are shown. The advantages of such equipment compared with those using anode modulation of the last hf stage are stressed.

**621.396.615.142.2:621.396.619.13** 2370  
Experiments on the Modulation of a Reflex Klystron—J. van Mulders. (*HF* (Brussels), no. 6, pp. 153-163; 1950.) Theory of the relation between the oscillation frequency of a reflex klystron and the reflector voltage is reviewed and measurements of the modulation index of an American klystron, Type 707B, are described. The modulation index was determined from photographs of the modulation spectrum displayed on a cro. The modulation distortion can at once be found from the modulation index when the static characteristic of the klystron (frequency as a function of reflector voltage) is known.

**621.396.619.23** 2371  
Rectifier Modulators with Frequency-Selective Terminations—D. G. Tucker. (*Proc. IEE* (London) vol. 97, pp. 205-207; May, 1950.) Discussion on 260 of February.

**621.396.619.23** 2372  
Non-Linear Distortion in a Cowan Modulator—Belevitch. (*Wireless Eng.*, vol. 27, pp. 164-165; May, 1950. Author's reply to 1552 of July (Tucker and Jaynes).

**621.396.61+621.396.621** 2373  
Radio Receivers and Transmitters [Book Review]—Amos and Kellaway. (See 2336.)

## TUBES AND THERMIONICS

**621.385:621.318.572** 2374  
An Electron-Beam Decimal-Counter Tube—D. L. Hollway. (*Nature* (London), vol. 165, pp. 856-857; May 27, 1950.) The tube described is 2.9 cm in diameter and 15 cm long, and operates at the relatively low anode voltage of 500 v with a beam current of 300  $\mu$ a and a frequency range 0-68 kc. The beam is caused to move in a path comprising alternate radial and circumferential steps by means of a special 5-plate deflector system with appropriate connections to a 5-segment slotted collector electrode. Visual indication of the count is given by projection on to the fluorescent screen of numbers cut in the back segments of the collector electrode.

**621.385.029.63/.64** 2375  
Wave Propagation in a Slipping Stream of Electrons: Small Amplitude Theory—G. G. Macfarlane and H. G. Hay. (*Proc. Phys. Soc.*, vol. 63, pp. 409-427; June 1, 1950.) Amplifying waves are found to travel along a slipping stream of electrons for all frequencies, a slipping stream being defined as one in which the electrons move in parallel paths with velocities varying with distance transverse to their motion. A slipping-stream tube may have the properties of a two-beam tube or a traveling-wave tube. The characteristics of one or the

other are displayed according to whether  $(V_a - V_{-a})/(V_a + V_{-a}) \leq 0.42$  the velocity of the electrons varying linearly from  $V_{-a}$  to  $V_a$  across the stream. A slipping stream inside a waveguide, capable of guiding a TM wave of slow phase-velocity in the absence of the electrons, acts very much as a traveling-wave tube with a uniform electron beam.

**621.385.029.63/.64** 2376  
Experimental Investigation of a Long Electron Beam in an Axial Magnetic Field—J. S. A. Tomner. (*Chalmers Tekn. Högsk. Handl.*, no. 92, 16 pp; 1950. In English.) An extension of an experimental investigation (561 of 1949) of the effect of the focusing magnetic field on the amplification of a traveling-wave tube. The form of the magnetic field produced by two helical focusing coils is studied and the collector current measured as a function of the number of turns of the coils. To avoid helix current due to space-charge effect for a helix of length 35 cm and diameter 3 mm, using a beam current of 1.5 ma and an accelerating voltage of 2,200 v, a field of 400 gauss or more is necessary.

**621.385.029.63/.64** 2377  
Traveling-Wave Tubes: Part 2—J. R. Pierce. (*Bell Sys. Tech. Jour.*, vol. 29, pp. 189-250; April, 1950.) The second installment (Chapters 4, 5, and 6) of a forthcoming book. Chapter discusses the two methods of analyzing iterated waveguide structures used in traveling-wave tubes, involving field theory and lumped-circuit analogues respectively. These methods are illustrated by analysis of typical simple structures. Chapter 5 is a discussion of group and phase velocities, gain and bandwidth, and the criteria by which the quality of a circuit should be judged. The helix and various resonator circuits are compared. In Chapter 6 the circuits are described in terms of normal modes of wave propagation. Part 1: 1810 of August.

**621.385.029.63/.64** 2378  
1000-Watt Traveling-Wave Tube—S. E. Webber. (*Electronics*, vol. 23, pp. 100-103; June, 1950.) An article based on a 1949 National Electronics Conference paper. An illustrated description and construction details are given of a water-cooled tube which as a power amplifier produces a power gain of about 25 at 450 Mc. The efficiency is 20 per cent when operated with a beam voltage of 5,000 and beam current of 1 amp. Unwanted oscillations are suppressed by an attenuator consisting of a conducting coating on the outside of the tube. The inherent bandwidth is more than adequate for most commercial applications where a wide transmission band is required. See also *Proc. NEC*, vol. 5, pp. 493-499; 1949.

**621.385.032.213.2** 2379  
Change of Mutual Conductance with Frequency—W. Raudorf. (*Wireless Eng.*, vol. 27, p. 164; May, 1950.) Author's reply to 1812 of August (Eisenstein). See also 1301 of June.

**621.385.032.216** 2380  
Contribution to the Study of Oxide Cathodes—F. Violet and J. Riethmüller. (*Le Vide*, vol. 4, pp. 687-720; November, 1949.) Reprint. See 2956 of 1949.

**621.385.15** 2381  
Secondary-Emission Valve—G. Diemer and J. L. H. Jonker. (*Wireless Eng.*, vol. 27, pp. 137-143; May, 1950. Correction, *ibid.*, vol. 27, p. 194; June, 1950.) "An experimental secondary-emission tube is described. The high figure of merit ( $gm/C=3.0$  ma/v-pF) that is obtained by adding one stage of secondary emission to an earthed-grid triode of rather conventional construction makes the tube useful as a wideband amplifier for those cases where a very low noise figure is not required; typical figures are: at 1 m wavelength 30 db gain  $G$  with a bandwidth  $B$  of 3.5 Mc, at 50 cm  $G=15$  db

with  $B=20$  Mc, at 30 cm  $G=10$  db with  $B=10$  Mc. The maximum power output is for  $\lambda>1$  m about 1.5 watts. At 7 m wavelength the noise figure amounts to 12 db; this rather high value is due to secondary emission. It is shown that for this secondary-emission noise a kind of space-charge smoothing effect exists."

621.385.18.001.4 2382  
Measurement of Tube Voltage Drop in Hot-Cathode Gas Tubes—Smith. (See 2291.)

621.385.2 2383  
Space Charge in Planar Diodes—C. S. Bull. (Proc. IEE (London), vol. 97, pp. 159-165; May, 1950.) Following Langmuir's assumptions, the author considers five variables: total emissivity  $i_s$ , anode current  $i$ , anode voltage  $V$ , the value  $V_m$  of Langmuir's "potential minimum" and the cathode-anode separation  $x$ . Of these only three are independent. From them are formed thirty partial differential coefficients which are tabulated. By using these, quantities not accessible to direct measurement can be calculated. Particular attention is paid to the "motional transconductance"  $(\partial i/\partial x) V_{is}$ , which is related to scale changes in tubes and to their microphonic properties; a method for determining this quantity is described

621.385.3:621.396.615.14 2384  
Electron Transit Time—M. R. Gavin. (Wireless Eng., vol. 27, p. 164; May, 1950.) Comment on 1820 of August (Chatterjee and Sreekantan).

621.385.832 2385  
Improved Production Methods for Television C.R. Tubes—(Engineer (London), vol. 189, pp. 482-483; April 21, 1950.) Description of an automatic machine installed at the Mul-lard factory for joining the necks and heads of cr tubes; the continuous production rate is 60 bulbs per hour.

621.385.832:535.371.07:621.396.9 2386  
Visibility on Cathode-Ray Tube Screens: Positive vs Negative Signals on an Intensity Modulated Scope—M. W. Harriman and S. B. Williams. (Jour. Opt. Soc. Amer., vol. 40, pp. 102-104; February, 1950.)

621.396.615.141.2:621.385.833 2387  
Electron-Optical Mapping of the Space-Charge Field in a Magnetron—(See 2315.)

621.385:[621.396.621+621.396.645 2388  
Electronic Valves: Books II-IV [Book Review]—Cleaver-Hume Press, London and N. V. Philips, Eindhoven, Holland.

Book II. Data and Circuits of Receiver and Amplifier Valves. 409 pp., 21s.

Book III. First supplement to Book II. 213 pp., 12s. 6d.

Book IV. Application of the Electronic Valve in Radio Receivers and Amplifiers. 416 pp., 35s. (Wireless Eng., vol. 27, pp. 193-194; June, 1950.) Book I was noted in 519 of March. Books II and III chiefly consist of details of the characteristics of Philips' tubes, one dealing mainly with the E, C, and K series and the other with the more recent E20, U20, D20, and U series and their applications. Book IV is in the nature of a wireless receiver textbook and is of general application. "The treatment throughout is very thorough but . . . almost entirely from the viewpoint of the designer of broadcast receivers."

621.385.83 2389  
The Theory and Design of Electron Beams [Book Review]—J. R. Pierce. Publisher: D. Van Nostrand, New York, 1949, 197 pp., \$3.50. (Wireless Eng., vol. 27, p. 166; May, 1950; Electronics, vol. 23, pp. 134-136; June, 1950.) "Written on the graduate level, [the book] discusses that material on electron beams which lends itself most readily to mathematical analysis, although some attention is also given to experimental techniques . . . [It] is clearly intended for those concerned with the formation and focusing of electron beams for use in such devices as low-frequency amplifiers, oscillators, and, especially, microwave tubes. . . . Comparisons between theory and practice are frequent. All chapters are followed by excellent problems. . . ."

#### MISCELLANEOUS

6:061.4 2390  
British Scientific Instruments exhibited at Olympia—(Metallurgia (Manchester), vol. 41,

pp. 401-413; May, 1950.) Short descriptions of equipment shown on the stand of the Scientific Instrument Manufacturers' Association at the 1950 British Industries Fair, including a wide range of electronic apparatus, magnetic and electrical instruments, electromechanical equipment, temperature measurement and control apparatus, optical equipment and vacuum plant.

621.39 Heaviside 2391  
The Heaviside Centenary—(Electrician, vol. 144, pp. 1747-1748; May 26, 1950.) Report of Heaviside Centenary Commemorative Meetings, May 1950. The opening paper, "Heaviside, the Man," by G. Lee, was followed by short tributes from leading scientists. Abstracts of papers on particular aspects of Heaviside's work will be found in the relevant sections.

621.39 Heaviside 2392  
Oliver Heaviside and his Layer—E. V. Appleton. (Wireless World, vol. 56, pp. 187-188; May, 1950.) A short appreciation of "Heaviside's work" with special reference to his suggestion concerning the possible existence of a conducting layer in the upper atmosphere. See also 2391 above.

621.3 2393  
Leitfaden der Elektrotechnik. Vol. 1: Grundlagen der Elektrotechnik [Book Review]—Moeller and Wolff. Publisher: B. G. Teubner Verlagsgesellschaft, Leipzig, 4th edn., 358 pp. (Wireless Eng., vol. 27, p. 193; June, 1950.) This textbook is unreservedly recommended to both students and teachers with a knowledge of German; the foundations of electrical engineering are laid with great thoroughness. A new chapter is included on the calculation of single-phase and three-phase transmission lines.

621.38/.39 2394  
Electronic Engineering Master Index for 1947-1948 [Book Review]—Publisher: Electronics Research Publishing Co., New York, 1950, 339 pp., \$19.50. (Jour. Frank. Inst., vol. 249, p. 340; April, 1950.) The second supplement of this reference work, which covers all aspects of the subject, and indexes material from about 250 scientific periodicals.

## Books (continued)

Applications of the Electronic Valve in Radio Receivers and Amplifiers. By H. G. Dammers, Ing. J. Haantjes, J. Otte, In. H. Van Suchtelen.

Published (1950) by Philips Technical Library, Eindhoven, Holland. 410 pages+7-page index+256 figures. 9×6½.

This book is one of a series of three from the Philips organization of Holland, one of the world's leading technical centers in lamps and electronics. These three volumes will bear the same title and the chapters will be numbered consecutively. Only this first volume has been printed, the other two being in preparation. This book contains chapters on rf and if amplification, mixing and oscillator circuits, capacitor tracking, parasitic effects, and distortion with the final chapter on detection. With this rather limited scope, it is apparent that in a large book, such as this one (over 400 pages), the subjects are covered in considerable detail. The treatment of each subject is from a fundamental, and, in most cases, mathematical viewpoint, with little or no mention of specific tube types.

It will probably be of limited interest to engineers in the U.S.A. engaged in the design of radio receivers, because there is no inclusion of material on FM circuits nor the circuits used in modern television receivers. This would appear to be true also of the two additional volumes in preparation.

The material in each chapter is well organized under subheadings, but the general index is quite limited in its coverage. There are references at the end of each chapter.

Engineers needing the basic mathematical relationship for the subjects covered will find this volume an excellent one for reference. The three volumes are to be made available also in the Dutch, German, and French languages.

W. C. WHITE  
General Electric Research Laboratory  
Schenectady, N. Y.

#### NEW PUBLICATIONS

The Geiger-Muller Counter, National Bureau of Standards Circular 490, 25 large double-column pages, illustrated, is avail-

able from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C., at 20 cents a copy. Remittances from foreign countries should be in United States exchange, and should include an additional sum of one-third the publication price to cover the cost of mailing.

The nature, construction, and use of the Geiger-Muller counter, one of the most important of present-day detectors of radioactive radiation, is concisely presented on an elementary level in this new booklet just published by the National Bureau of Standards.

In addition to the treatment of the Geiger-Muller counter itself, the booklet discusses methods of detecting counter pulses, applications of counters to quantitative measurements, proportional counters, and the preparation and filling of Geiger-Muller counters. Also included are a number of examples of special forms of counters which have been developed and a discussion of some of the electronic circuits commonly used to obtain an indication of the response of the counter to radiation. A bibliography of scientific papers is presented.